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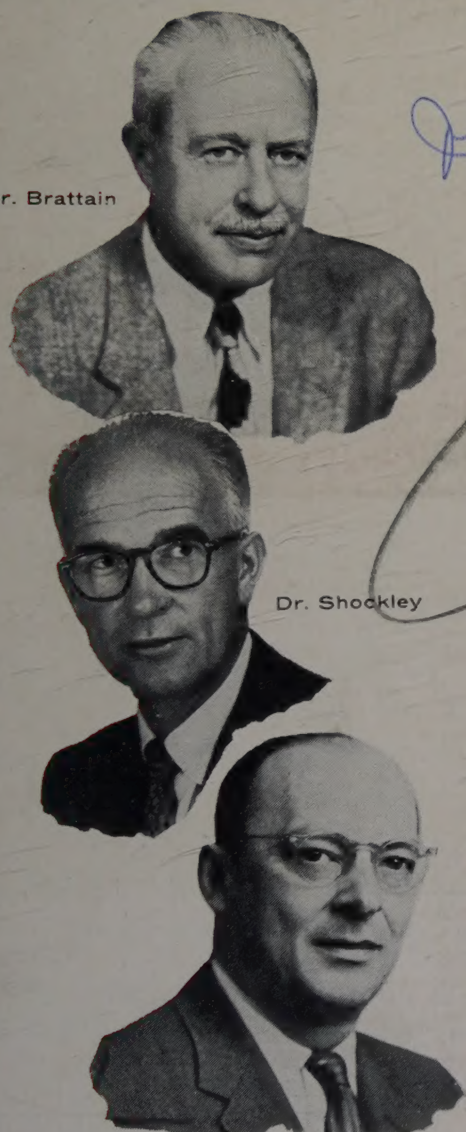
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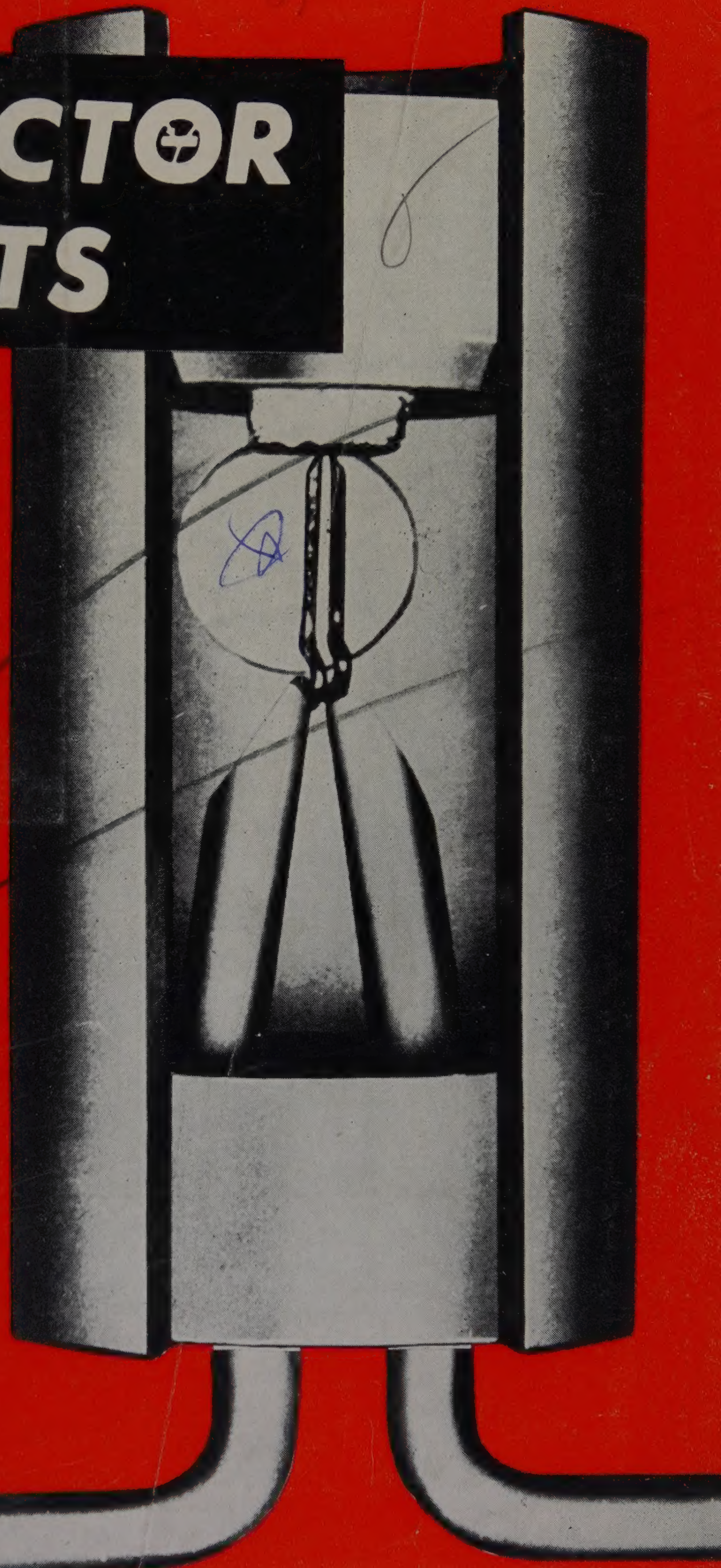
Dr. Bardeen



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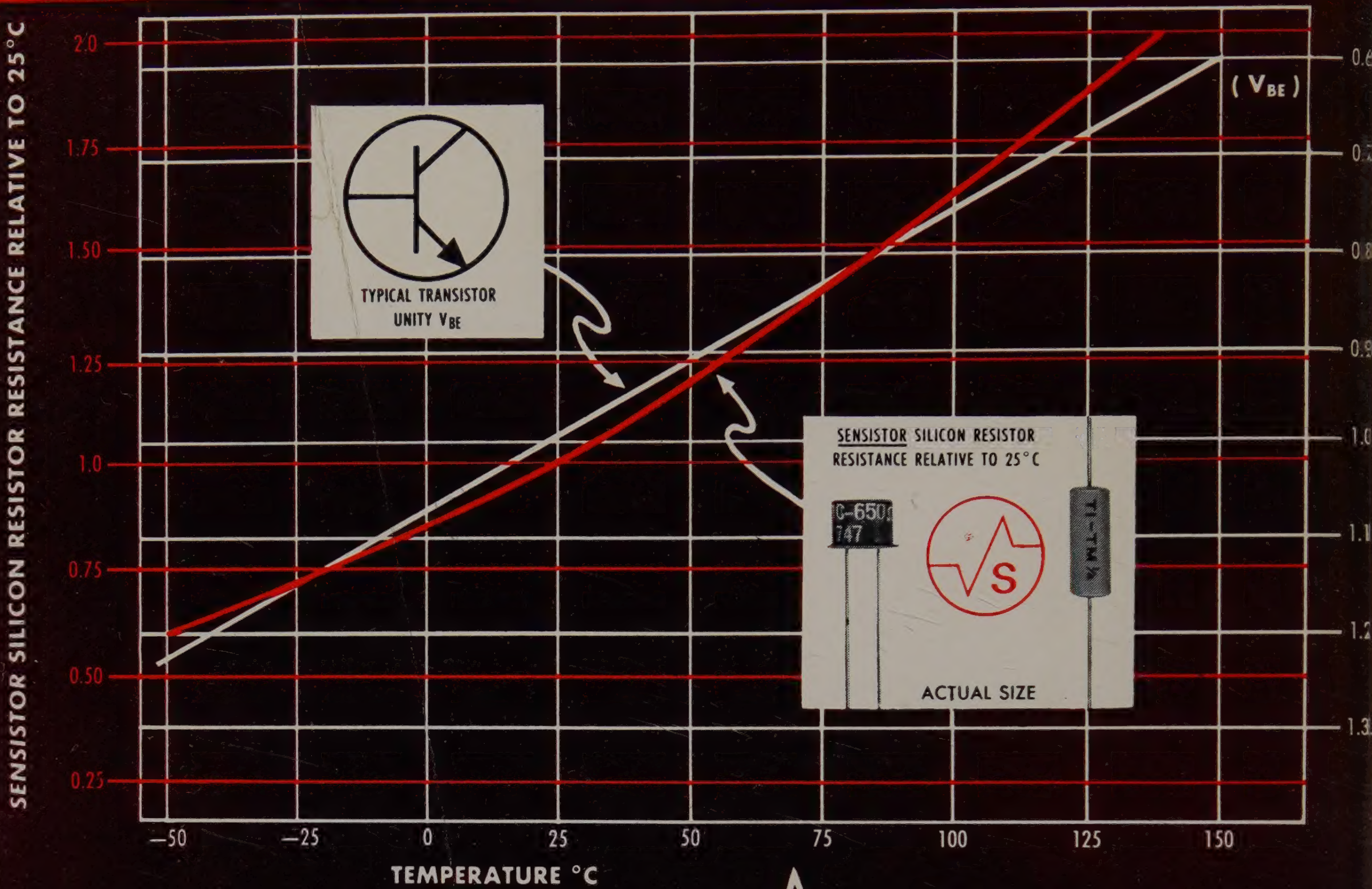

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SANFORD R. COWAN, Publisher

May/June 1958 Vol. 1, No. 3

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Front Cover

The front cover this month shows the original point contact transistor announced by Bell Telephone Laboratories on July 1, 1948 and the three men responsible for the discovery of the transistor. Dr. Shockley is presently the Director of the Shockley Semiconductor Laboratory, a division of Beckman Instruments, Inc. Dr. Brattain is now engaged in research connected with the physics of semiconductor surfaces at Bell Telephone Laboratories, and Dr. Bardeen is with the Department of Physics at the University of Illinois, Urbana, Illinois. We, at Semiconductor Products Magazine, want to take this opportunity to congratulate Drs. Shockley, Bardeen and Brattain on this, the 10 Anniversary of the Transistor, and to wish them many more happy and fruitful years.

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Editorial . . .

Transistor's Tenth Anniversary

Ten years ago, on June 30, 1948, the first demonstration of the Transistor was made by Bell Telephone Laboratories. The key investigations which brought the transistor to reality were carried out by Dr. John Bardeen and Dr. Walter H. Brattain. The general research program leading to the transistor was initiated and directed by Dr. William Shockley. Since that time these three scientists were honored with the Nobel Prize for this work and the industry has grown to a point where, for the year 1957, sales of semiconductor diodes and transistors reached a 142 million dollar total. The climb in such sales has been a steady one and according to universal opinion will continue to be steady, reaching the billion dollar mark between 1960 and 1970. It must be emphasized that these figures pertain only to the devices themselves, and not to the complementary components which, together with these devices, make up various equipments. Great strides have been made in these ten years in perfecting materials, devices and manufacturing processes, and human ingenuity is proceeding at a rapid rate towards applying these devices in ever-widening fields. In addition, the advent of semiconductors has given rise to parallel development in related component development, particularly in the realm of miniaturization, where transistors are unique in their application.

Tremendous strides have been made from the point of view of basic materials not only in manufacturing germanium and silicon with smaller impurity percentages but in augmenting these intrinsic materials with an increasing number of intermetallic compounds with properties that increase the temperature, frequency, and power handling capabilities of semiconductor devices. Intrinsic silicon is now made with an impurity content of less than 1 part in 6 billion parts of silicon. Silicon carbide devices are now operating at temperatures of over 500° C.

Many new devices are constantly being announced reflecting the trend towards higher power and higher frequencies. The frequency frontier is now well over 1000 *mc*. The power handling capabilities of transistors extends into the KW regions. *R-F* power handling capabilities is approaching 1 watt at 100 *mc*. The switching speeds of silicon transistors now extend to within the 30 to 50 millimicrosecond range. We begin to hear more and more of devices like 4-Layer Diodes, Microalloy Diffused Base Transistors, Transistor Tetrodes, Dynistors, Trinistors, Triode Diffused Base Tetrodes, Spacistors, Unijunction Transistors, Silicon Controlled Rectifiers, Shift Register Transistors, Sensistors (a semiconductor resistance device), and others which space restrictions do not allow us to mention.

From the point of view of applications, both present and imminent, an impressive array presents itself. These include computer information processing and switching, AM and FM radios, car radios, transistorization of test equipment, power supplies, transistorized *d-c* to *d-c* inverters, power transmission applications, two-way communications systems, TV receivers, automobile electrical and fuel injection systems, controls for machinery, and atomic radiation meters.

The promise and glamor of this new industry constantly invites enterprising individuals to enter the field of manufacture. The industry is young and isn't sewed up by a far cry. Of those undertaking the risk, many will fail and some will succeed. The primary factors entering into the success of such an undertaking are competent personnel and good planning.

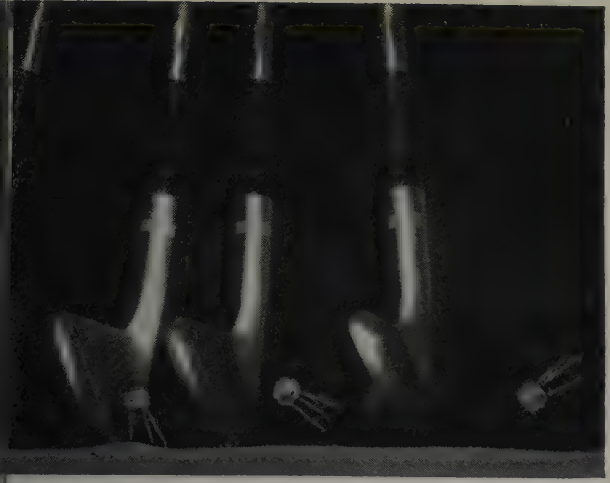
The future of this industry is sparked by the promise of an untold number of materials yet to be discovered, devices to be developed, manufacturing processes to be invented and applications to be exploited. We may conclude by pointing out that all areas of human endeavor can be, and eventually will be, served by semiconductor devices in one form or another.

Giving Credit where Credit is Due

Our charts on currently-announced diodes and rectifiers, and transistors are meeting with favorable comment from the industry. We attribute the success of these charts in a great degree to the amount of informational items we are able to include within the format of the physical dimensions of the magazine. Giving credit where credit is due, we wish to thank the following individuals for their technical help in designing the chart layouts: Richard Keller—General Electric, Robert Hammann—Sylvania, Marian Magargal—Lansdale Tube (Philco), Ted Finger—General Transistor, Bernard Reich—U.S. Army Signal Corps, Henry Tulchin—Derivation & Tabulation Associates (DATA), Jack Gillette—Raytheon, Bob Sollinger, Jr. — General Electric, Paul Petrack — I. T. & T., Walter Bonner—I. T. & T., Steve Klevans—Automatic Mfg. Div. (General Instrument), Ed Vandeven—General Electric.

Samuel L. Marshall

remendous impact and vibration



CLUB TEST" General Electric transistors with Fixed-Bed Mounting en struck full force with a No. 2 Iron. After traveling forty yards, tests they still worked perfectly.

"JACKHAMMER TEST" Another G-E transistor with Fixed-Bed Mounting was taped to a pneumatic drill, which was then operated for ten minutes. When the transistor was removed, tests showed it still worked perfectly.

Ceramic disk guards against major causes of transistor failure

neral Electric's new Fixed-Bed Mounting, critical ele- of the transistor are welded flat on a disk of ceramic. any impact must be great enough to damage the disk before transistor failure can occur. In conventional ds of manufacture, impact need only penetrate the stor's metal case in order to damage the standard ht header.

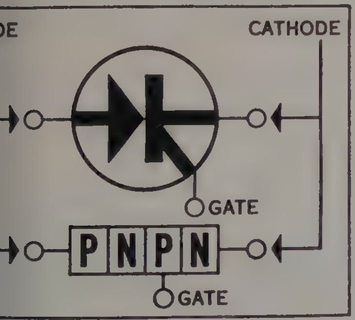
cause of their many suspended parts, standard upright ers are also subject to inertial stress at a number of s. General Electric's Fixed-Bed Mounting eliminated *it one* of those parts—the suspended aluminum emitter And this is provided with enough slack to absorb al stress, with connection points so securely welded the unit withstands far more than the military cen- ge test of 20,000 G's.

To eliminate thermal stress, the coefficient of expansion of G.E.'s ceramic disk has been made equal to that of the semiconductor metal. Previously, enough "play" had to be allowed to absorb alternate expansions and contractions, thereby reducing the strength and stability of the unit.

The Fixed-Bed Mounting's electrical elements lie flat, in close contact to the transistor case, providing greater heat conduction out through the case. Therefore, the fixed-bed construction cuts down junction temperature, making it possible to double the power dissipation of the same transistor made with upright-header construction.

Fixed-Bed Mounted units have exceeded all standard shock, centrifuge and temperature-cycling tests. General Electric's unijunction transistor (see below) now has this feature.

ew G-E Controlled Rectifier rectifies and ntrols current up to 5 amperes at 300 v.



ntrolled rectifier is a four-layer silicon with a "gate" to which a signal can be l to control forward current. It can handle han one kw of power.

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General Electric's new silicon controlled rectifier acts like a thyratron. In the reverse direction, it's a standard rectifier. But it will also block forward current until either a critical breakover voltage is exceeded or a signal is applied to the third lead. Then it switches to a conducting state and acts as a forward-biased silicon rectifier.

The controlled rectifier can be actuated by a little as 15 mw. Breakdown occurs at speeds approaching a microsecond, after which voltage across the device is so low that current is determined by the load. This enables the user to control a large anode-to-cathode current with an extremely small amount of power, or to switch power from high impedance to low impedance in microseconds.

Applications include replacement of relays, thyratrons, magnetic amplifiers, power transistors and conventional rectifiers. Sample quantities of the controlled rectifier are now available. Prices will be sent on request.

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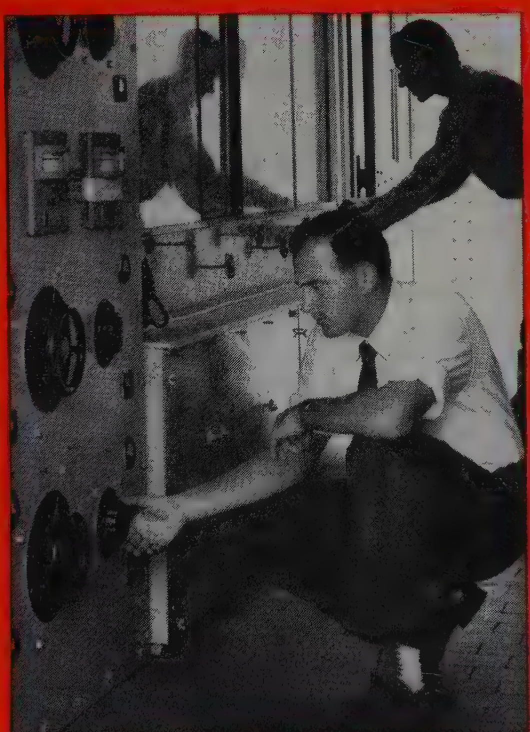

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At present, the polycrystalline material contains a small concentration of a Group V element which segregates rapidly in zone refining. No other elements, such as tantalum, gold, zinc, iron, manganese, molybdenum, potassium, sodium, bismuth, and cobalt, appear to be present even when tested by the most sensitive analytical technics such as activation analysis.

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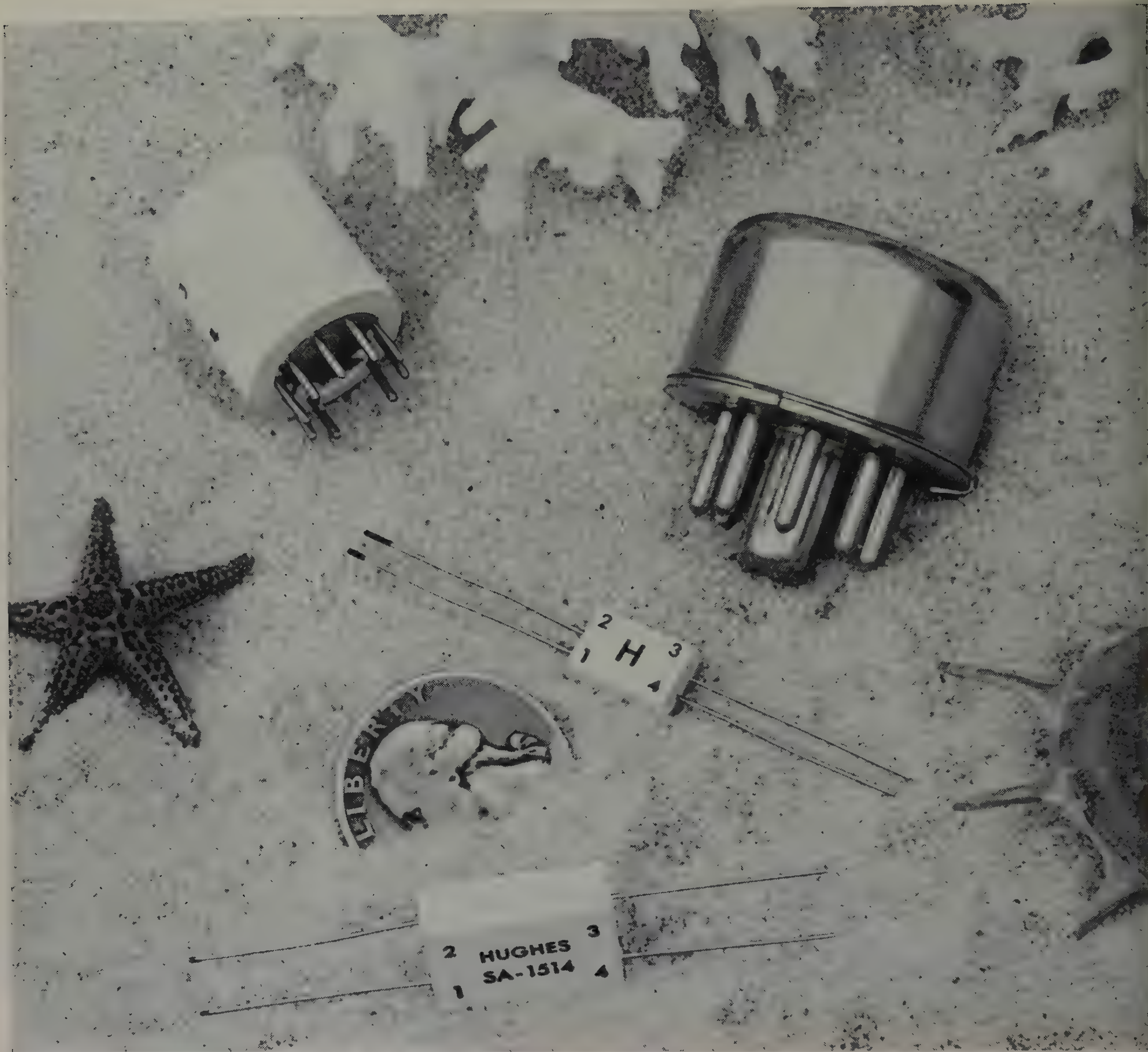

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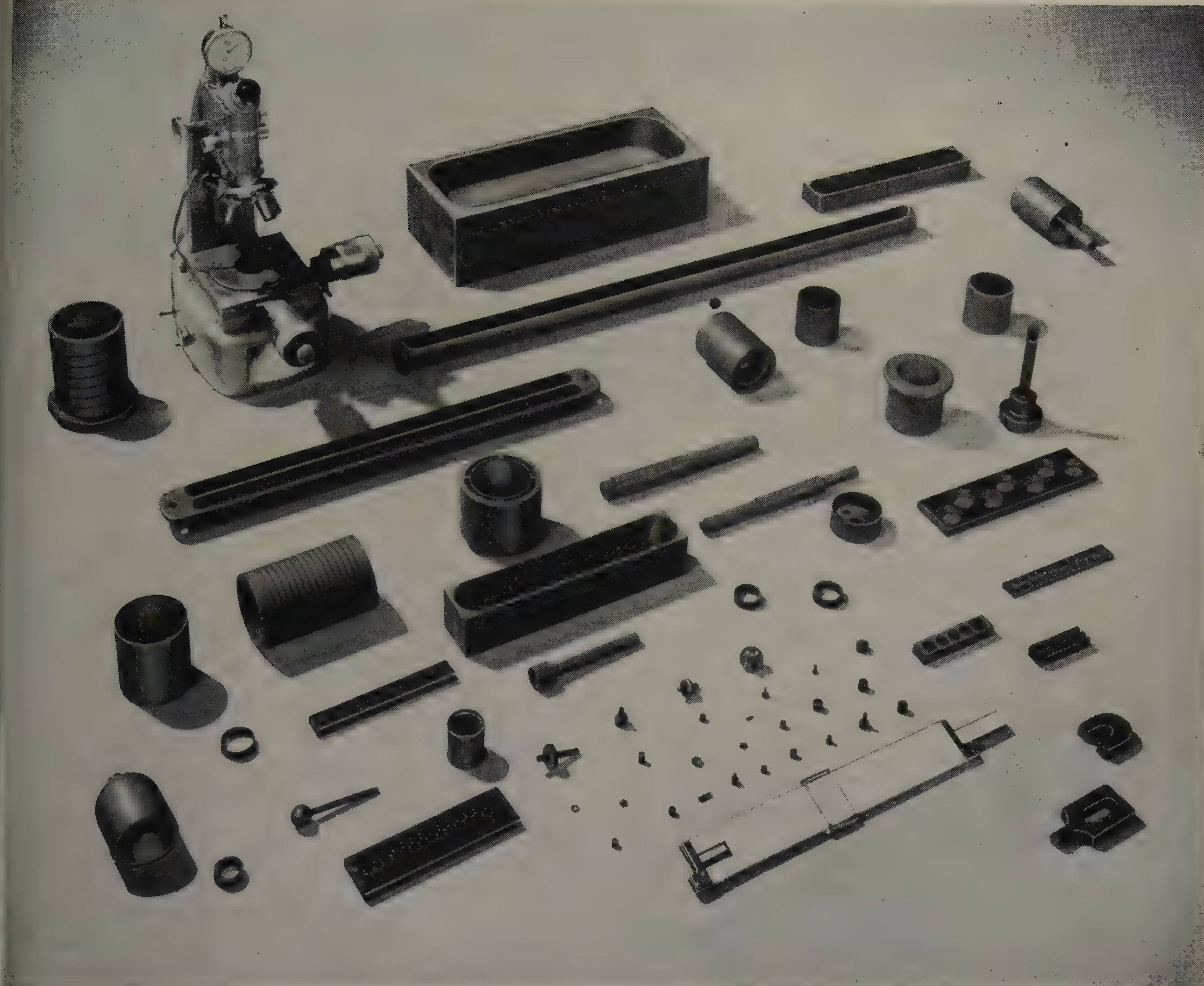

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Hole Storage Delay Time and Its Prediction

C. D. SIMMONS*

Transistors are often operated in the saturated condition in switching circuits because of the low series resistance, low dissipation, and immunity from transistor parameter variation this type of operation affords. There is a charge storage phenomena connected with this type of operation which results in a finite time delay when the transistor is turned "off". It is the purpose of this article to describe and define this effect and then to derive a single criteria for the prediction of this time delay. Experimental data is then presented and the utility of the method discussed.

SATURATION may be defined using the circuit of *Fig. 1*. As the base current, I_B, is increased from zero, the collector current will increase until the voltage drop across the load resistor approaches the battery voltage. At this point, the collector to emitter voltage approaches zero. Further increases in base current will not increase the collector current materially. When this condition is reached, the base-to-emitter voltage is larger than the collector-to-emitter voltage so that the base is more negative than the collector (*p-n-p* transistor is assumed). This means that the collector-to-base diode becomes forward biased as the collector current nears its limit value. The transistor is said to be in saturation or "bottomed" when it is operated such that the collector-to-base diode is forward biased. A simpler, but less exact, definition is that a transistor is in saturation if more current is caused to flow in the base than is required to maintain the collector current at the value determined by the circuit. Since the collector voltage will be almost zero, the saturated collector current will be approximately:

$$I_{CS} \approx \frac{V_{CC}}{R_L}. \tag{1}$$

The base current which would be required to just maintain this collector current is

$$I_{BS} = \frac{I_{CS}}{\beta_0}, \tag{2}$$

where β_0 is the grounded emitter current gain "at the edge of saturation," which means that it is the current gain when the collector voltage is such that the collector-to-base voltage is zero. Since, in saturation, the base current is greater than the value given by (2), we may define an excess base current as:

$$I_{BX} = I_B - I_{BS}. \tag{3}$$

If the transistor of *Fig. 1* is provided with some excess base current and if the base drive conditions are then changed so that the transistor would no longer be operated in the saturated condition, it will be observed that the collector current will not fall immediately. For example, if a scope is connected to the collector of a saturated transistor and then the base circuit is abruptly opened, the delay effect shown in *Fig. 2* will be observed. The time delay between the opening of the base circuit and the start of the fall (10% down) of the collector current is defined as the hole storage delay time, t_s. It will be found that this time is a function of: (*a*) the excess base current which was provided prior to turn-off, (*b*) the way in which the transistor is turned off, and (*c*) the geometry of the transistor used. Because of the many advantages gained in switching circuits by using the transistor in its saturated mode, it is desirable to be able to predict the storage time of a given transistor in a particular circuit.

Ebers and Moll[1] have analyzed an ideal transistor and Moll[2] has derived an equation for the hole storage time under a number of simplifying assumption which include the assumption that the transistor is linear and non-variant. These assumptions permit a simplified mathematical treatment of the problem without seriously affecting the accuracy of the result. Moll has pointed out that since the transistor is assumed completely linear, the *saturated* transistor may be considered as two parallel unsaturated transistors connected back to back as shown in *Fig. 3*.

We have already pointed out that the collector is forward biased when the transistor is saturated, so that it will act as an emitter as well as a collector. The

*Chief Applications Engineer, Lansdale Tube Co. Division of Philco Corp., Landsdale, Pa.

[1]J. J. Ebers and J. L. Moll, "Large-Signal Behavior of Junction Transistors," *Proc. of the IRE,* Dec. 1954.

[2]John L. Moll, "Large Signal Transient Response of Junction Transistors," *Proc. of the IRE,* Dec. 1954.

emitter, which is forward biased, will also collect as well as emit. Using *Fig. 3,* we will identify I_{EF} as the current injected at the emitter, I_{CR} as the current collected at the collector, I_{CF} as the current injected at the collector, and I_{ER} as the current collected at the emitter (these symbols are Moll's). Since the external currents are determined by the circuitry when the transistor is operated in saturation, the following equations must hold:

$$I_C = I_{CR} - I_{CF}, \tag{4}$$

$$I_E = I_{EF} - I_{ER}, \tag{5}$$

$$I_E = I_C + I_B. \tag{6}$$

Since we have assumed the transistor to be completely linear, we may also write

$$I_{CR} = \alpha_F \, I_{EF}, \tag{7}$$

$$I_{ER} = \alpha_R \, I_{CF}, \tag{8}$$

where α_F is the normal grounded-base, low-frequency current gain of the transistor (in the forward direction) and α_R is the grounded-base current gain in the reverse direction (inverted connections). Each of these α's has an alpha cutoff frequency associated with it (frequency at which α has dropped 3 *db*) which we shall designate $f\alpha_F$ and $f\alpha_R$ respectively.

Moll has shown that, in the common emitter connection, the hole storage time is given as

$$t_S = \frac{f_{\alpha F} + f_{\alpha R}}{2\pi \, f_{\alpha F} \, f_{\alpha R} \, (1 - \alpha_F \alpha_R)} \ln \frac{I_{B1} - I_{B2}}{I_{BS} - I_{B2}}, \tag{9}$$

where I_{B1} is the base current prior to turn-off, I_{BS} is given by *Equation* (2), I_{B2} is the base current during the turn-off which may be positive or negative, and where the other terms have already been defined. This equation holds for the "current turn-off" case where turn-off is affected by changing the current from I_{B_1} *to* I_{B_2}. I_{B_2} will be negative if reverse base current is supplied to turn the transistor off. The constant in *Equation* (9) is completely determined by the transistor geometry, whereas the value of the logarithm is almost completely dependent upon the circuit (only I_{BS} is influenced by a transistor parameter). As given, the constant requires the measurement of four separate transistor properties for its evaluation. It can be shown that this constant may be obtained from one simple measurement if all of the assumptions used in the derivation of (9) hold.

In fact it is found (see Appendix) that

$$\frac{f_{\alpha F} + f_{\alpha R}}{2\pi \, f_{\alpha F} \, f_{\alpha R} \, (1 - \alpha_F \alpha_R)} = \frac{Q_{SX}}{I_{BX}}, \tag{10}$$

where Q_{SX} is the excess stored charge in the base, *i.e.* the total amount of charge which must be removed to take the unit out of saturation, and where I_{BX} is given by *Eq.* (3). Q_{SX}/I_{BX} is a constant, since all of the terms on the left hand side of (10) are assumed to be constant. Let us define this ratio as the hole storage factor:

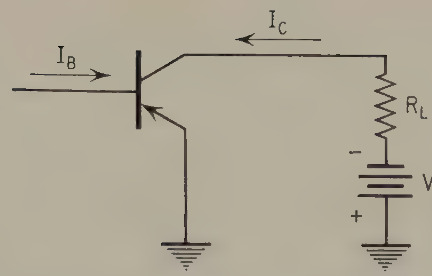

Fig. 1—Reference circuit for defining saturation.

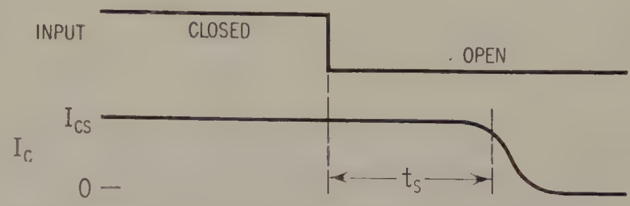

Fig. 2—Scope presentation illustrating the delay in current drop when transistor operation changes abruptly from a saturated to an unsaturated condition.

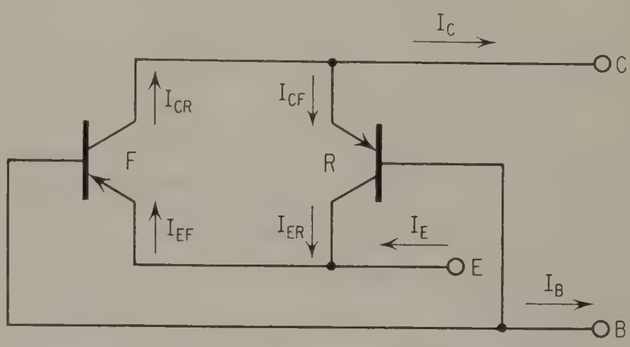

Fig. 3—Representation of a saturated transistor by two unsaturated transistors placed back to back.

$$K_S = \frac{Q_{SX}}{I_{BX}}. \tag{11}$$

Then *Equation* (9) becomes:

$$t_S = K_S \ln \frac{I_{B1} - I_{B2}}{I_{BS} - I_{B2}}. \tag{12}$$

To evaluate K_S it is now merely necessary to measure the stored charge per unit excess base current. This may be done using the circuit of *Fig. 4*.

Consider the conditions before the positive pulse is applied to the base. There is no *d-c* path in the collector; hence, no *d-c* current can flow in the collector. The emitter current, which is adjusted by R_1, will all pass through the base terminal. From *Equation* (3) the excess base current will be equal to the emitter current. R_2 is a very small resistor so that there is negligible drop across it. The positive pulse is of sufficient amplitude so that the vacuum diode, D_1, is turned off during the pulse interval. Since the base is positive during this interval, the stored charges in the base will now exit through the collector and charge

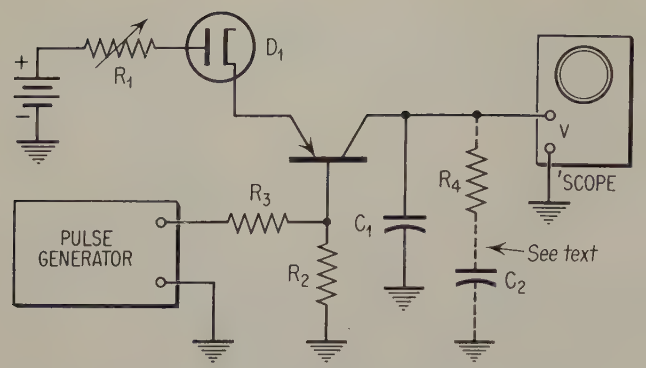

Fig. 4—Circuit for evaluating hole storage factor.

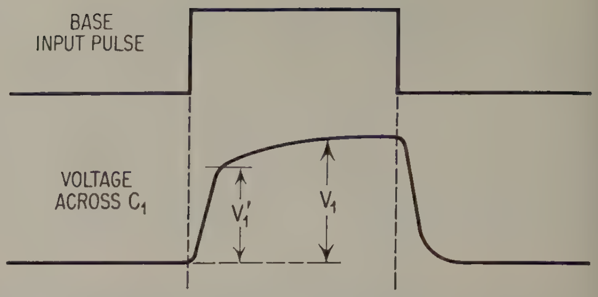

Fig. 5—Scope presentation resulting from application of pulse indicated in test set-up of Figure 4.

C_1. This will continue until all of the stored charges are exhausted after which C_1 can no longer charge. The scope presentation will be as shown in *Figure* 5. The upward slope to the wave form is caused by two effects. Part of the slope is due to the back leakage current (I_{CO}) of the collector diode. The pulse voltage fed to the base of the transistor is kept small (usually 3*V*) to minimize this contribution. The other effect is due to the geometry of the transistor. The cross section of a typical transistor is shown schematically in *Fig. 6*. Most of the charge will be stored in *Region A* of the base and may exit quickly. A small part of the charge is stored in *Region B*. This charge must travel a longer path to be removed and hence, these charges exit more slowly. This effect becomes important in very fast switching transistors where the actual junction areas are small.

The change in voltage, V_1, is a measure of the total stored charge while the voltage V_1', is a measure of the charge stored in the active area (*Region A*). In a high speed switching circuit only the charge in the active area exhibits itself as hole storage delay time. The remaining charge, which exits more slowly, shows up as a tail on the fall time. In slower circuits, where the hole storage delay time is long enough to allow the charges in *Region B* to exit, all of the charge contributes to the hole storage delay time.

If we concern ourselves only with the total charge for the moment, we may say that:

$$V_1 = \frac{Q_S}{C_1}. \tag{13}$$

All of this charge is excess stored charge since, by *Equation* (32), the stored charge at the "edge of saturation" is zero (in this circuit). Hence

$$K_S = \frac{Q_{SX}}{I_{BX}} = \frac{C_1 V_1}{I_E} \tag{14}$$

where I_E is the emitter current in the circuit of *Fig. 4* between base pulses. This K_S is the number which is to be inserted in *Equation* (12) when dealing with slower speed applications. In very high speed work we may define:

$$K_S' = \frac{C_1 V_1'}{I_E}. \tag{15}$$

This number is to be used in *Equation* (12) when calculating the hole storage delay time in these applications. In very high speed transistors the difference between K_S and K_S' is about 30%. The difference is considerably less in lower speed units. The measurement of K_S' can be simplified by connecting a differentiating network across C_1 as shown in dotted lines in *Fig. 4*. The time constant is arranged so that the voltage across C_1 peaks at the value, V_1'.

There are two special cases of *Equation* (12) which are of importance. If I_{B2} is zero we may write:

$$t_S = K_S \ln \frac{I_{B1}}{I_{BS}}. \tag{16}$$

We may define a circuit gain as

$$\beta_C = \frac{I_{CS}}{I_{B1}}, \tag{17}$$

which is the ratio of saturated collector current to supplied base current. Substituting this and (2) in (16) we have

$$t_S = K_S \ln \frac{\beta_o}{\beta_C}, \tag{18}$$

which is the storage time encountered when the transistor is turned off by opening the base. *Equation* (12) may be rewritten using (3) as:

$$t_S = K_S \ln\left(1 + \frac{I_{BX}}{I_{BS} - I_{B2}}\right). \tag{19}$$

If the turn-off current is a large reverse current such that

$$| I_{B2} | >> I_{BS}$$

and

$$| I_{B2} | >> I_{BX}$$

then the storage time is given approximately as:

$$t_S \approx K_S \frac{I_{BX}}{| I_{B2} |}. \tag{20}$$

Equations (10) and (11) shows that K_S depends upon the α and $f\alpha$ values which, although treated as constants in our discussion, are actually slightly dependent upon operating point in an actual transistor. Hence, the precaution should be taken of measuring K_S with the base current at approximately the level to be expected during actual circuit operation.

Equation (19) is plotted in *Fig. 7* on semi-log paper. One may calculate the base current factor which is the horizontal axis of that curve, enter the curve, and read off the ratio of hole storage time to K_S. Also shown on that curve are actual measurements that were taken on a large junction audio frequency transistor at a number of collector current and base current levels. The *X's* refer to measurements on one unit and the *o's* refer the measurements on another unit. In general, it has been found that the hole storage time on low and medium frequency units may be predicted reliably to within about 20% by this theory and the storage time of high frequency units to within about 30%. The typical error is considerably less than the limits just given.

APPENDIX:

Derivation of Eq. (10) We shall first derive expressions for I_{CF} and I_{EF} in the saturated condition. Since we have assumed that the parameters are constant, the current gain will be non-variant so that:

$$\beta_o = \frac{\alpha_F}{1 - \alpha_F} \tag{21}$$

Substituting this in (3) we have, in saturation:

$$I_B = I_{B1} = I_{BX} + I_C \frac{1 - \alpha_F}{\alpha_F} = I_{BX} - I_C + \frac{I_C}{\alpha_F} \tag{22}$$

From (5), (8), and (22) we have:

$$I_E = I_C + I_B$$

$$I_{EF} - I_{ER} = I_C + I_{BX} - I_C + I_C/\alpha_F$$

$$I_{EF} - \alpha_R I_{CF} = I_{BX} + I_{EF} - I_{CF}/\alpha_F$$

The I_{EF}'s cancel and we are left with:

$$-\alpha_R I_{CF} = I_{BX} - I_{CF}/\alpha_F$$

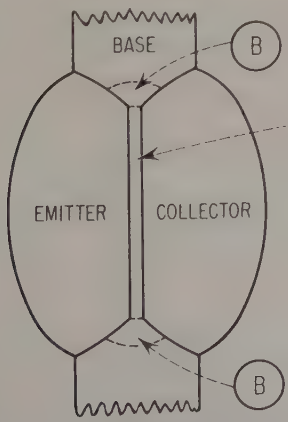

Fig. 6—Storage areas in a typical transistor.

$$I_{CF} = I_{BX} \frac{\alpha_F}{1 - \alpha_R\alpha_F} \tag{23}$$

Combining (4), (7), and (23):

$$I_{EF} = \frac{I_{CR}}{\alpha_F} = \frac{I_C}{\alpha_F} + \frac{I_{CF}}{\alpha_F} = \frac{I_C}{\alpha_F} + I_{BX} \frac{1}{1 - \alpha_R\alpha_F} \tag{24}$$

We have assumed in the foregoing that the transistor is in a steady state prior to turn-off time. Moll has indicated the necessary considerations for the case where the transistor has not been in the saturated condition for a sufficient length of time to establish these equilibrium conditions. The length of time required to establish equilibrium is approximately equal to the unsaturated rise time of the transistor.

A schematic representation of the carrier density within the saturated transistor is shown in *Fig. 8*. The lines represent plots of carrier density versus distance through the base. In an unsaturated transistor the density is zero at the collector since all carriers arriving at the collector are instantly collected. The line labeled n_F gives the carrier density associated with the normal forward transistor, *F*, of *Fig. 3*. The line n_R gives the density associated with the inverted transistor, *R*, of *Fig. 3*. The total of these two gives the actual carrier density function in the transistor. Since this carrier distribution must be set up in the transistor in order to maintain the steady state saturated condition, we may say that the saturated operation requires the storage of a certain amount of charge. If the operating conditions of the transistor are to be changed, then this charge carrier density must be changed. The dotted curve of *Fig. 8* is the carrier density when the transistor is operated at the edge of saturation. The shaded area of *Fig. 8* represents the charge which must be removed in order to take the transistor out of saturation. The name, "hole storage delay time," comes from the fact that it is the time required to remove this charge. Since

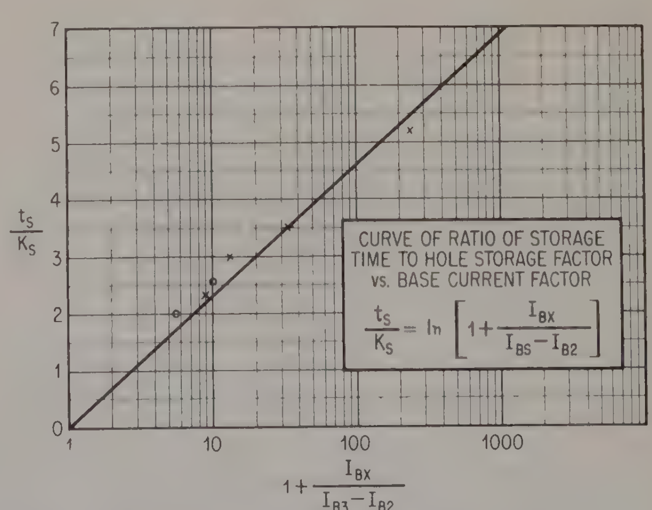

Fig. 7—t_S/K_S plotted against the base current factor.

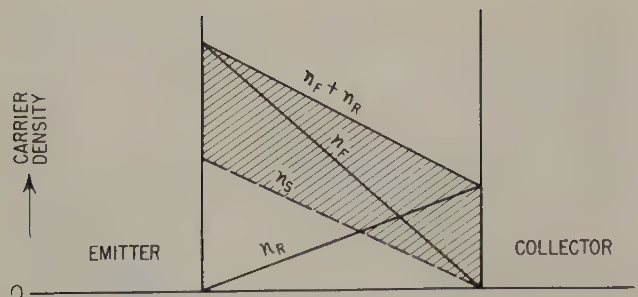

Fig. 8—Diagrammatic representation of the carrier density within a saturated transistor.

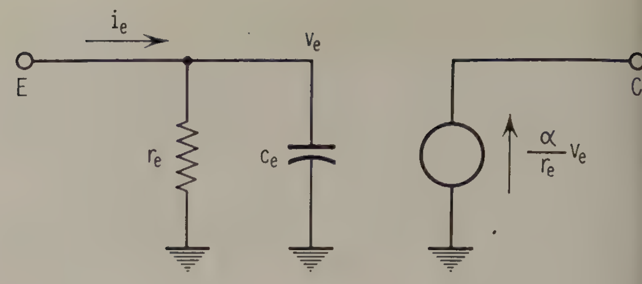

Fig. 9—Equivalent circuit representation of linear transistor with current drive.

this time is proportional to the quantity stored, it follows, therefore, that the frequency response of the transistor also is a function of the stored charge.

This charge storage may be thought of as due to a diffusion capacitance. Granting all of Moll's assumptions, we may represent the linear transistor with current drive as shown in *Fig. 9*. Note that the assumption has been made that there is no feedback. The voltage, v_e, will be given by

$$v_e = i_e \frac{r_e}{1 + j\,\omega\, c_e\, r_e} \tag{25}$$

where ω is the radian frequency of the current, i_e. The collector current will be given by:

$$i_C = \frac{\alpha}{r_e} v_e = \frac{\alpha}{1 + j\,\omega\, c_e\, r_e} i_e. \tag{26}$$

If we write *Equation* (26) in Laplace transform we have

$$I_C(s) = \frac{\alpha}{1 + s\, r_e\, c_e} I_E(s), \tag{27}$$

which is the equation used in Moll's paper. The C_e of these equations is the diffusion capacitance as seen at the emitter. *Equation* (26) shows that the current gain will be down 3 *db* when

$$f_\alpha = \frac{1}{2\pi\, r_e\, c_e}, \tag{28}$$

hence, *Equation* (28) gives the alpha cutoff frequency of the transistor. If the emitter current of *Fig. 9* is a steady state *d-c* current, then the charge stored in the diffusion capacitance will be:

$$Q = C_e\, V_e = c_e\, r_e\, I_E.$$

Using *Equation* (28)

$$Q = I_E \frac{1}{2\pi\, f_\alpha}, \tag{29}$$

where I_E is the steady state emitter current, and $f\alpha$ is the alpha cutoff frequency of the transistor.

We have already shown that the saturated transistor may be considered to be two linear, non-saturated transistors connected back-to-back which may each be described by *Fig. 9*. The total charge stored will equal the sum of the charges stored in each one. Hence, the total stored charge in saturation is:

$$Q_S = I_{EF} \frac{1}{2\pi\, f_{\alpha F}} + I_{CF} \frac{1}{2\pi f_{\alpha R}}. \tag{30}$$

At the "edge of saturation," I_{CF} is equal to zero and the emitter current will be equal to (see *Equation* (7)):

$$I_E = \frac{I_C}{\alpha_F}. \tag{31}$$

Hence, the total stored charge at the edge of saturation is:

$$Q_o = I_E \frac{1}{2\pi\, f_{\alpha F}} = \frac{I_C}{2\pi\, \alpha_F\, f_{\alpha F}}. \tag{32}$$

The total amount of charge which must be removed to take the unit out of saturation, or the excess stored charge will be given by:

$$Q_{SX} = Q_S - Q_o = \frac{1}{2\pi\, f_{\alpha F}} \left(I_{EF} - \frac{I_C}{\alpha_F} \right) + \frac{I_{CF}}{2\pi\, f_{\alpha R}}. \tag{33}$$

We have calculated the value of I_{CF} and I_{EF} in *Equations* (23) and (24). If we substitute these in (33), we have:

$$Q_{SX} = I_{BX} \left[\frac{f_{\alpha R} + \alpha_F\, f_{\alpha F}}{2\pi\, f_{\alpha F}\, f_{\alpha R}\, (1 - \alpha_F\, \alpha_R)} \right].$$

If we assume that α_F is approximately equal to unity, then

$$Q_{SX} = I_{BX} \frac{f_{\alpha R} + f_{\alpha F}}{2\pi\, f_{\alpha F}\, f_{\alpha R}\, (1 - \alpha_F\, \alpha_R)},$$

or:

$$\frac{f_{\alpha R} + f_{\alpha F}}{2\pi\, f_{\alpha F}\, f_{\alpha R}\, (1 - \alpha_F\, \alpha_R)} = \frac{Q_{SX}}{I_{BX}} \tag{34}$$

This factor is the device constant of Moll's equation (*Equation* 9) for hole storage time.

High-Current Switching Applications of Low-Power Transistors*

C. HUANG‡, W. F. PALMER‡, C. M. CHANG‡

There is a need for switching transistors capable of controlling some hundreds of milliamperes of current while exhibiting good transient response. Since the high storage time and low frequency response of conventional power transistors renders them unsuitable, we are led to consider the use of low-power types at low duty factors. Measurements on characteristics of importance in high-current applications show that switching operation of transistors intended for low-current, low power uses is permissible where high peak currents are involved; provided that attention is paid to limiting the maximum junction temperature.

THERE IS a need for switching transistors capable of controlling some hundreds of milliamperes of current while exhibiting good transient response. Since the high storage time and low frequency response of most conventional power transistors renders them unsuitable, we are led to consider the use of the low-power types, even though certain ratings appear prohibitively low.

Frequently the duty factor of high-power dissipation in these applications can be held sufficiently low that we will have no difficulty with a low dissipation rating, but the current rating (on the order of 5 to 10 *ma* for most useful low-power types) definitely prohibits their use at currents one or two orders of magnitudes greater.

Let us review the more important factors which should be considered before concluding that these low-power transistors may be operated at high peak currents and dissipations.

While dissipation and voltage ratings may be readily set, it is more difficult to determine a realistic maximum of emitter or collector current. One simple method is, of course, to define arbitrarily the maximum permissible collector current as (*1*) that which is permissible or useful in Class A signal applications or (*2*) that at which the current gain B (large signal) has fallen to the minimum useful value and even this is arbitrary since we have now brought circuit considerations into the problem—not device life alone. It is, however, reasonable to do this, since devices are required to operate in circuits which are, in turn, required to perform some desired operation.

Based on a paper presented at the 1956 Wescon, Los Angeles, Calif.

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These, at least, furnish us with a starting point, in particular the latter, *i.e.,* if this value of current gain occurs at current much in excess of the existing nominal maxima then we may attempt to answer the question, "is it really permissible to operate the transistor at such high values of current?"

Theoretical studies may indicate that such operation is harmless and a statistical life of some 100,000's of hours may be expected. Actual experience with such figures is, of course, lacking and we may have to extrapolate by a few orders of magnitude to obtain estimated life figures.

Life tests (limited in time to some thousands of hours) indicate that high-current operation of several commercial and experimental transistor types is not more injurious than within-rating operation since the time variations in such vital parameters as I_{co}, V_{cmax}, β, B (etc.) are essentially the same for shelf, rated, and high-current operation.

To be useful in such applications as are of interest to many of us, the transistor must have the following characteristics:

(a) The large-signal current gain, B, must be sufficiently high at the peak or maximum current required.

(b) Switching times must be sufficiently low to yield a good pulse shape (this is largely a matter of frequency response though a too-high storage time is undesirable because of the turn-off delay introduced).

(c) Rise and fall times must be sufficiently low that they do not (*1*) introduce an undesirably high component of average dissipation or (*2*) exceed the peak dissipation rating.

(d) Saturation voltage must be low enough that the average dissipation rating is not exceeded.

(e) Base voltage must also be low since base dissipation may be a significant fraction of the total dissipation.

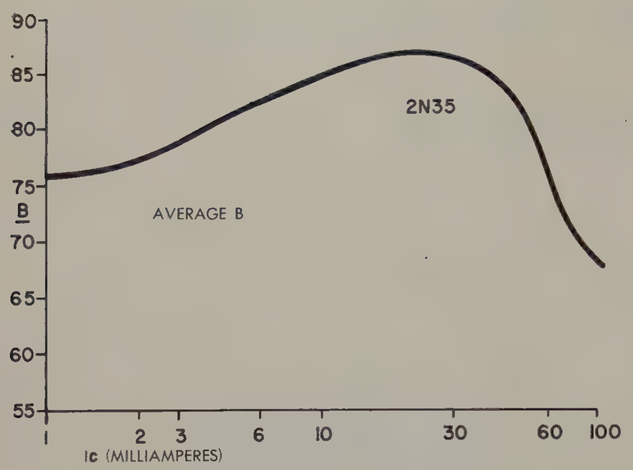


Fig. 1—B vs I_c 2N35

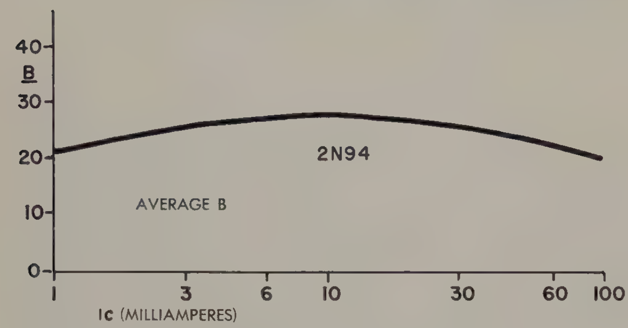


Fig. 2A—B vs I_c 2N94

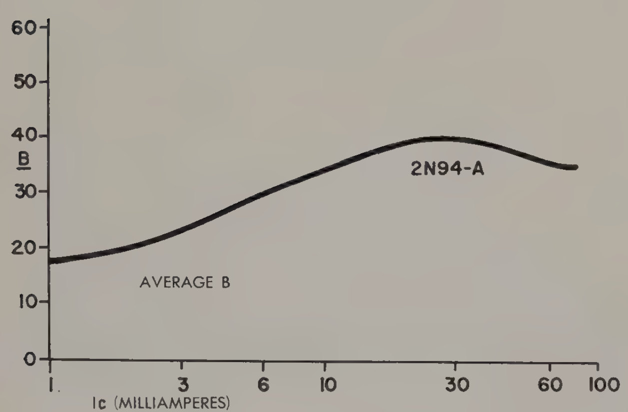


Fig. 2B—B vs I_c 2N94A

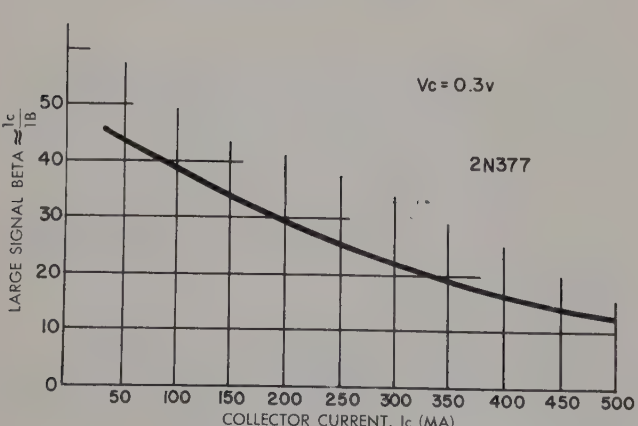


Fig. 3—B vs I_c 2N377

(f) I_{co} and I_{eo} must be reasonably low.

We may have to consider all of these factors individually or as combinations—and in terms of the proposed application (s).

Considering these factors individually and with respect to each other:

1. B, D-C Large Signal Current Gain

In terms of a particular application we may set limits of useful B at the particular peak current required—and determine whether or not a particular transistor type (s) will satisfy them. *Figs. 1, 2a, 2b,* and *3* show B *vs* I_c for typical 2N35, 2N94, and 2N377 types of transistors. Note that B is usefully high at some hundreds of *ma* in each case (the 2N94 and 2N94A types are approximately symmetrical with respect to normal and inverse beta).

2. Switching Times

Low frequency transistors having long storage times may be satisfactory where pulse rates of some tens of *kc/s* are required *i.e.,* where pulse widths are some tens of μsec. However, the use of high-frequency types is necessary when good pulse shapes are desired for pulse widths of less than 10 μsec. There are, of course, high-current applications for both types depending upon the pulse width and repetition rates required. *Figs. 4* to *17* show switching characteristics of typical 2N35, 2N94 and 2N377 transistors. *Fig. 4* indicates the measurement techniques employed in obtaining these data. Note that in all cases these transients are on the order of tenths of μsec to microseconds, depending upon the particular transistor type considered. Choice of a suitable transistor and drive conditions will allow us to obtain a good pulse shape.

3. Rise and Fall Times and Dissipation Considerations

It is important to know the effective value of the transistor's thermal time constant (die to can or package), particularly if we propose to switch the maximum volt-amperes, and the thermal time-constant is relatively short. *e.g.,* if a transistor has an effective time-constant of 1 msec, a temperature rise of 1°C per *mw*, and peak ratings of 25 *v* and 100 *ma*, then the instantaneous maximum dissipation during switching is over 0.6 *w*, and we may compute an initial temperature rise on the order of 0.5°C per *μsec*. If we may accomplish switching (on or off, rise or fall) in even 5 *μsec* (which is not difficult for low-frequency transistors), then junction temperature has increased about 2.5°C during this period. If we are within the average dissipation rating of the transistor (*i.e.,* suitably-low duty factors are used) we may ignore, as a first approximation, the contribution of these high instantaneous dissipation periods (during pulse rise and fall) to average dissipation. We may then proceed to compute average dissipation as the "on" dissipation—duty factor product. The contribu-

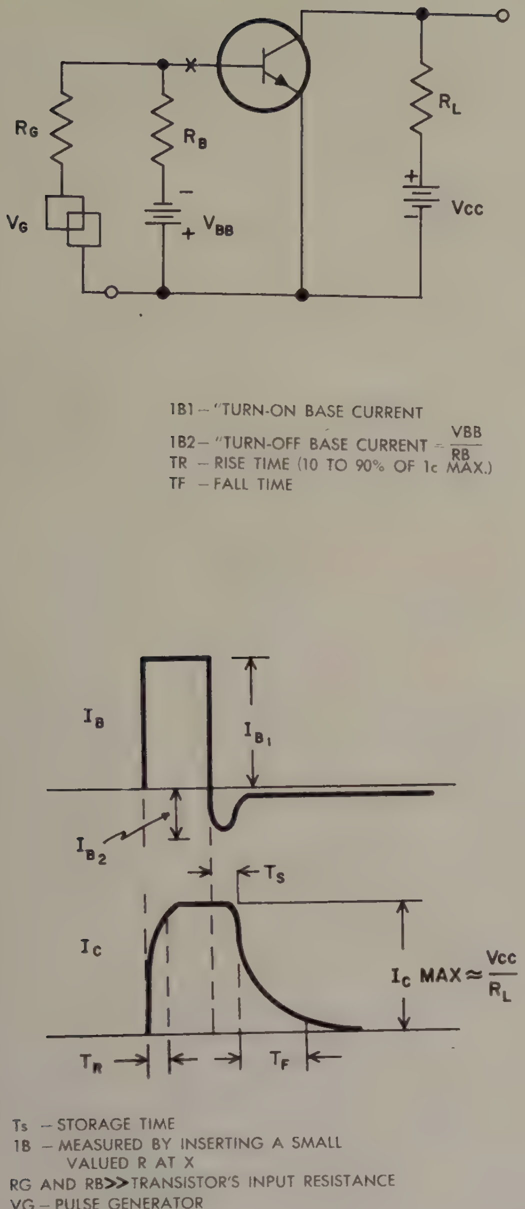


Fig. 4—Conditions of Measurement of Switching Transients.

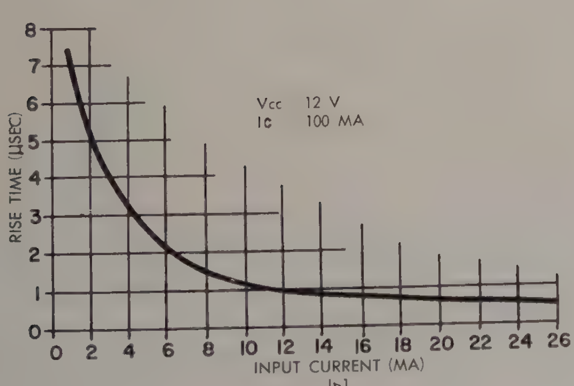


Fig. 5—Rise Time, 2N35

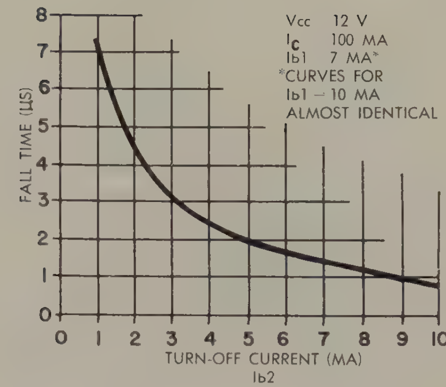


Fig. 6—Fall Time, 2N35

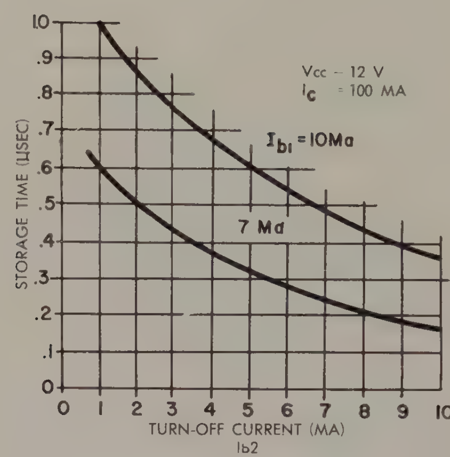


Fig. 7—Storage Time, 2N35

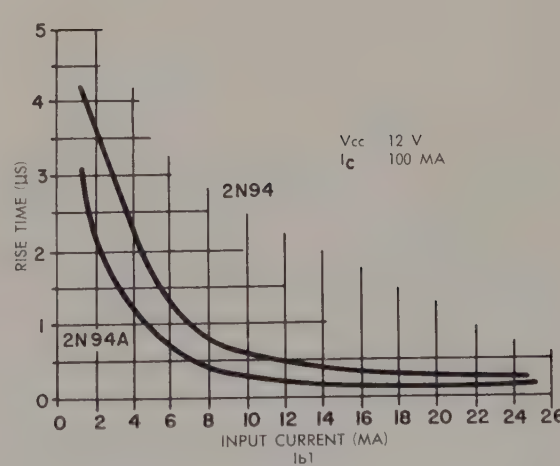


Fig. 8—Rise Time, 2N94 and A

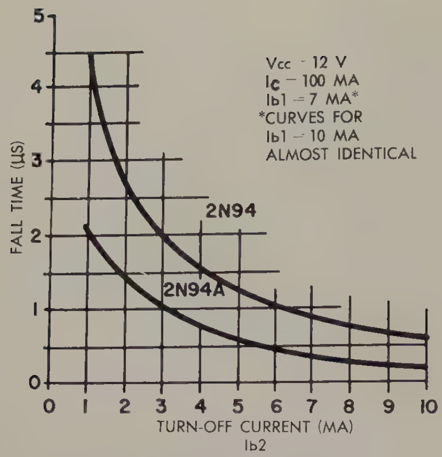


Fig. 9—Fall Time, 2N94 and A

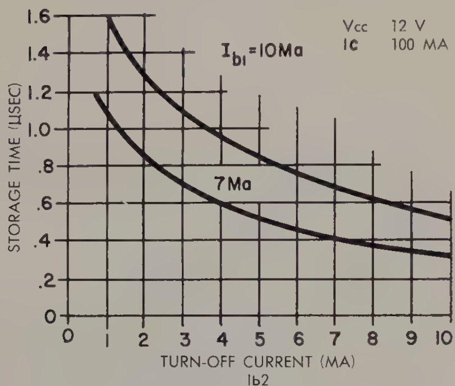


Fig. 10—Storage Time, 2N94

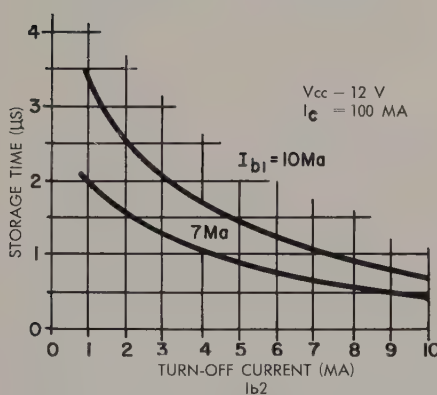


Fig. 11—Storage Time, 2N94A

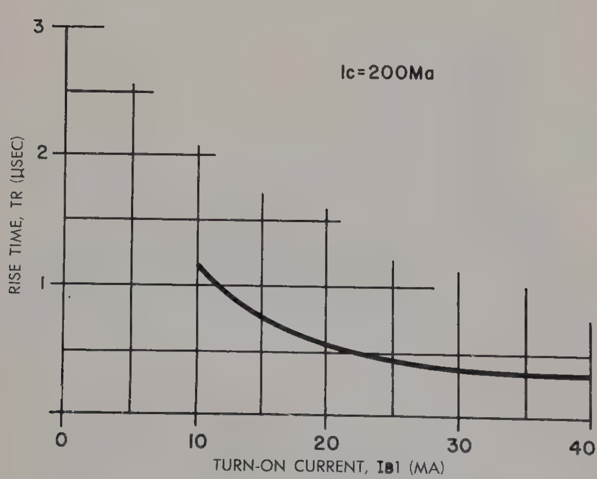


Fig. 12—Rise Time, I_c = 200 ma, 2N377

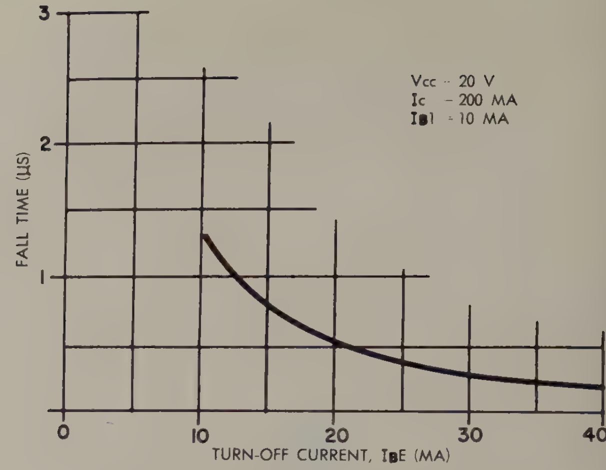


Fig. 13—Fall Time, I_c = 200 ma, 2N377

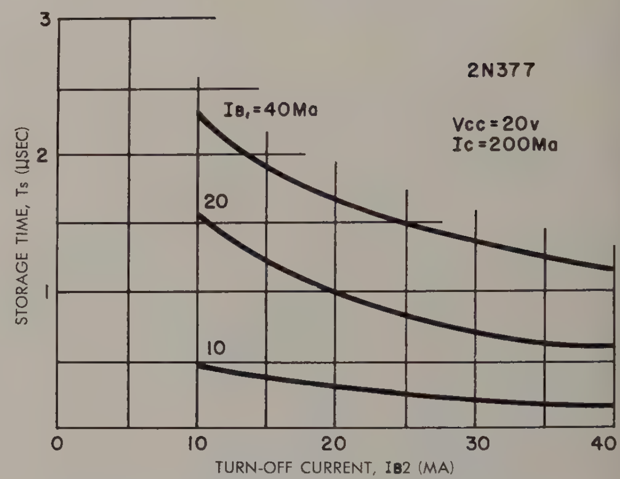


Fig. 14—Storage Time, I_c = 200 ma, 2N377

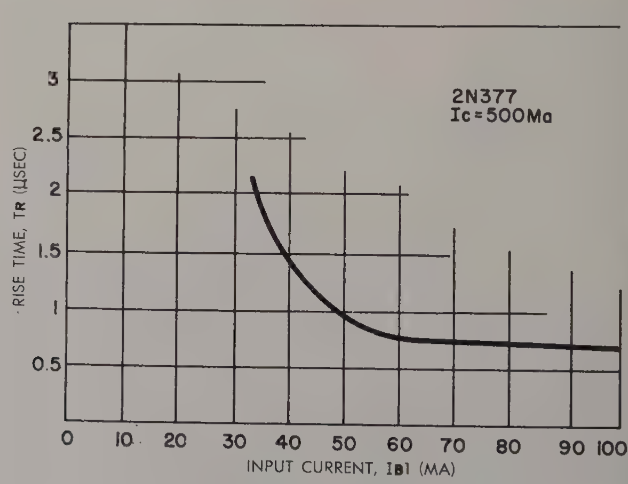


Fig. 15—Rise Time, I_c = 500 ma, 2N377

APPROXIMATE VALUES OF THERMAL TIME CONSTANTS

Type No.	P Max. Aver. (mw)	P Max. Inst. Pk. 10 μs Max. (W)	Vce Max. (E Biased Off) (V)	Ic Max. Cont. (ma)	Ic Max. Inst. Pk. (ma)	Thermal Time Const. (msec-) (Approx.)	T_J Max.	Cut-off Freq. (Typical) (mc/s)
N35	50	1	25	50	100	1	75° C	0.9
N214*	125	2.0	25	100	200	5	75° C	0.9
N94**	50	1	20	50	100	1	75° C	3.5
N94A**							75° C	6.0
N377**	150	2.5	25	200	500	10	85° C	5.0

*The 2N214 is a higher dissipation version of the 2N35 (other characteristics are similar).

**Though not indicated, these types are approximately symmetrical.

Note: The Thermal Time Constant referred to is junction-to-case.

tion of "off" dissipation may be similarly calculated and added where it is significant.

In the cases of the transistor types discussed below, this approximation is a valid one for most practical applications because of suitable combinations of characteristics.

Approximate values of thermal time constants for these transistors are included in the table above.

4. Saturation Voltage (at the peak current drawn)

The contribution of V_s to average dissipation is readily understood—in the case of the transistor which is "on" for a time not much-less-than its thermal time-constant, the "on" dissipation must not exceed the everage rated value. Where the "on" time is much lower, the transistor again serves as a "thermal integrator" and we may use the "on" dissipation—duty factor method of computing average dissipation (of also 5 below). *Fig. 18* shows the collector saturation voltage characteristics of the several transistor types mentioned above.

5. Base Voltage (V_B)

The base dissipation may introduce a significant component of "on" dissipation, but may usually be neglected as an approximation once a trial calculation is made. Typical values of V_b @ $I_b = 10\ ma$ are tabulated below:

2N35, 2N94, 2N94A ⟶ 0.55 V

2N377 ⟶ 0.80 V

6. I_{CO} and I_{EO}

In particular the former must be low enough to be negligible relative to the I_c to be drawn, and should (as a product with V_{cc}) contribute little to the average dissipation even at relatively high operating temperatures, or its contribution must be included in the average dissipation.

Transistors

Characteristics of the 2N35, 2N94, and 2N377 families of transistors are such as to permit relatively

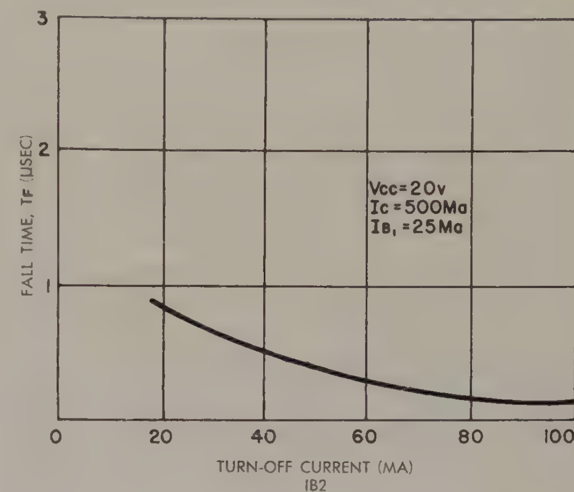


Fig. 16—Fall Time, $I_c = 500$ ma, 2N377

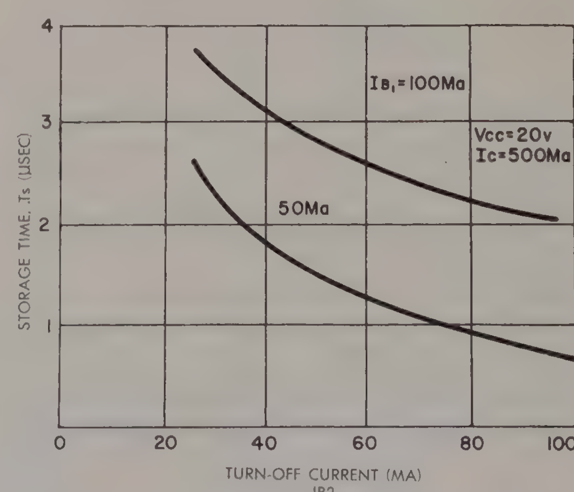


Fig. 17—Storage Time, $I_c = 500$ ma, 2N377

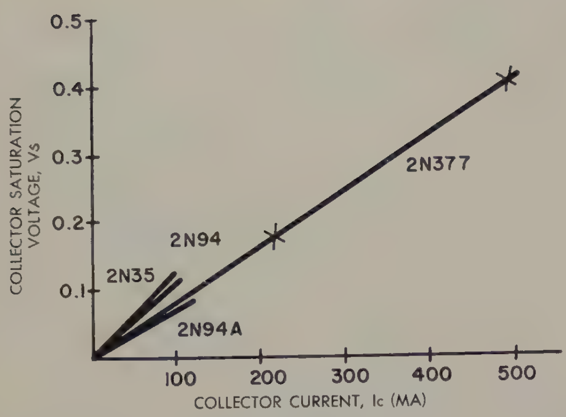


Fig. 18—Saturation Voltage Characteristics, 2N35, 2N94, 2N94A, 2N377

high-current operations of these types in appropriate circuits. Ratings on these three types for these applications are tabulated below.

Conclusion

Transistor characteristics of importance in high-current applications show that switching operation of transistors intended for low-current, low-power uses is permissible where high peak currents and dissipation are involved, provided that attention is paid to limiting the maximum junction temperature to that specified by the manufacturer. Examples have been given available types which are useful at peak currents of 100 to 500 *ma*.

Transistor Servo Amplifier Output Stages

ALLEN J. KEEGAN*

Power output stage of transistor servo amplifiers deserve extensive design effort since they have a large control over the volume, weight, and cost of the final package. Four types of common emitter output stages and their power supplies are analyzed with emphasis placed on selection of the optimum configuration for a given application.

TRANSISTORIZED amplifiers are enjoying wide use in servo-mechanisms for military applications. The electrical parameters of these amplifiers are dictated by the individual system but many have common requirements. Most of these amplifiers are used to drive a servo motor and its associated load for computation or positioning purposes. Installation compartments demand the smallest, lightest "black boxes" possible. Hence, packaging considerations cannot be ignored during the circuit design if a miniature, reliable, serviceable, low cost unit is to be realized.

Past experience in the package layout of servo amplifiers indicate the largest percentage of the total volume is occupied by the output stage. Considerable cost is concentrated in this area also. Thus, it is expedient to channel design effort toward improvement of the final stage. It must be borne in mind that the power supply electrical and physical requirements are dictated to a large extent by the output circuit.

Since these circuits are primarily used for power amplification, the common emitter configuration is most often used, rather than a common base or a common collector circuit, because of its more desirable power gain characteristics. This should not be construed to imply that these other two configurations have no place in servo amplifiers. Their features can be used to great advantage in some instances.

A single transistor biased for Class A linear operation can be used when low power is required by the motor load. This circuit has the advantage of simplicity, but it suffers in some other respects.

Common emitter Class A amplifiers are characterized by large quiescent collector currents. If the motor's tuned control phase is located directly in the collector circuit, motor performance will suffer because of iron saturation and heating caused by the *d-c* current. A coupling transformer can be used to elimi-

**John Oster Manufacturing Co.*

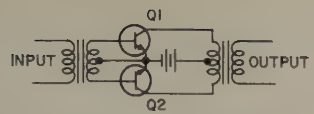


Fig. 1—Conventional Class B push pull stage

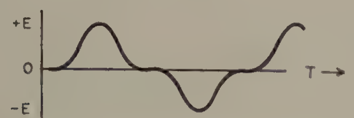


Fig. 3—Cross-over distortion

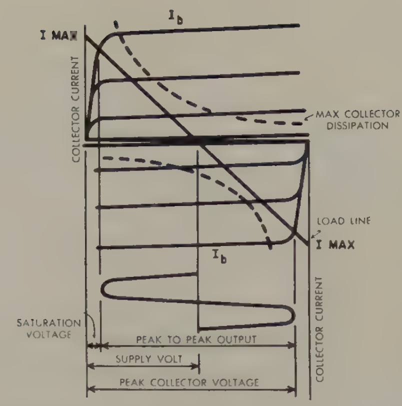


Fig. 2—Graphical presentation of Class B push pull operation

ate these effects. Collector circuit load can be optimized for maximum power output by selecting the proper transformer turns ratio. The primary open circuit inductive reactance at the operating frequency should be at least ten times the reflected resistance to minimize phase shift. This necessitates a large number of primary turns which aggravates the saturation problem created by the quiescent current. Therefore, the coupling transformer must consist of a large core stacked with a controlled airgap. Transformer efficiency will lower the total efficiency below the theoretical 50% of the ideal Class A circuit.

Some of the disadvantages of Class A power amplifiers can be overcome by the conventional Class B push-pull circuit shown in *Fig. 1*. During one-half cycle of input voltage Q_1 conducts collector current while Q_2 conducts during the other half cycle since the bases and emitters are at the same *d-c* potential. In the ideal case the *d-c* components of collector current of the transistors are equal, resulting in a cancellation of *d-c* magnetization of the transformer. Thus, for a given power level this center tapped transformer can be smaller than the one used in the single-ended circuit. Any desired resistance reflected into the primary can be obtained by adjusting the turns ratio.

The determination of the collector to collector resistance is augmented by a graphical analysis keeping in mind the limits of the transistors. Collector current, voltage and power dissipation ratings must not be exceeded. It is necessary to derate the power dissipation of the transistor if the amplifier will be operated at elevated temperatures. Reference to *Fig. 2* illustrates that the operation of each transistor is within all limiting parameters. Note that at full drive (maximum output) the collector current is less than the maximum rating and that the derated collector dissipation is not crossed by the load line.

It should be emphasized that the peak voltage appearing at the collector of Q_1 occurs when Q_2 is driven to full output. Hence, the sum of the collector supply voltage and the peak of the collector signal must not exceed the rated collector voltage. For transistors having a small collector saturation resistance, the supply voltage must be restricted to one-half of the rated collector voltage. The supply voltage can surpass this half-way point by a small amount if the transistors have a high saturation resistance.

Having once established the appropriate path of operation, it is possible to determine the available output power and the collector to collector *a-c* resistance (output transformer primary). The slope of the line represents the resistance seen by each collector independently. Because of the transformer action between halves of the transformer primary, the collector to collector resistance, is equal to four times the slope of the load line.

Output power can be calculated as follows:

$$P_o = \frac{E_{max}}{\sqrt{2}} \times \frac{I_{max}}{\sqrt{2}} = \frac{(E_{cc}\text{-}E_{sat})\ (I_{max}\text{-}I_{co})}{2}$$

where: P_o = Power output
E_{max} & I_{max} = Peak value of signal
E_{cc} = supply voltage
E_{sat} = Collector saturation voltage
I_{co} = Collector leakage current.

if E_{sat} & I_{co} are small

$$P_o = \frac{E_{cc}\ I_{max}}{2}$$

$$R_{cc} = \frac{E_{cc}}{I_{max}} \times 4 \qquad I_{max} = \frac{4\ E_{cc}}{R_{cc}}$$

where: R_{cc} = Collector to collector resistance

$$P_o = \frac{2\ E_{cc}^2}{R_{cc}}$$

As previously, the open circuit reactance of this

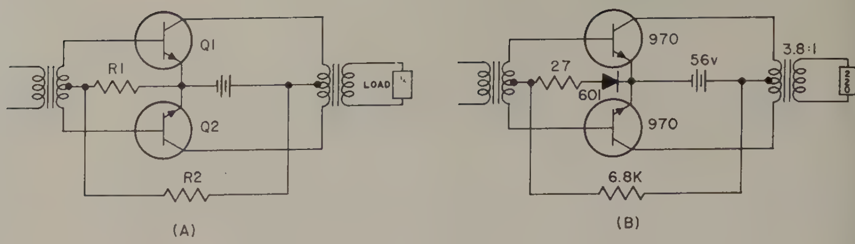


Fig. 4a & 4b—Forward bias to eliminate cross-over distortion

transformer should be a minimum of ten times the collector to collector resistance. Since it is impossible to maintain a perfect balance between collector currents, it is advisable to specify the expected unbalance (usually not more than 10%) to avoid transformer saturation.

Small signal output from the basic Class B push-pull amplifier shown in *Fig. 3* is badly distorted. This is referred to as cross-over distortion and is caused by the threshold nature of emitter junction characteristics. Cross-over distortion is more severe in stages using silicon transistors than those using germanium transistors because of the higher energy gap associated with silicon semiconductors. Cross-over distortion can be eliminated by forward biasing the emitter junction to the threshold point. For germanium transistors this will be approximately 0.1*v* while 0.6*v* is needed for silicon transistors. See *Fig. 4a.* The selection of R_1 is a result of a compromise between power supply current and gain.

If the forward bias is obtained from fixed resistors, Class AB operation will exist at elevated temperatures causing reduced output and excessive power supply current while Class C operation will prevail at reduced ambients. This performance impairment is a result of the temperature dependency of the emitter junction threshold voltage. Class B operation can be maintained throughout a wide ambient range by incorporating a temperature sensitive element in the biasing network. A logical choice of this element is a diode of the same material as the transistor, since the threshold voltage changes of both the diode and the emitter junctions will be matched. This eliminates any complex shaping networks associated with standard readily available thermistors. An additional small series resistor may be required to completely avoid cross-over distortion if the forward voltage drop of the diode is less than the amount needed by the transistor emitter junction. The circuit of *Fig. 4b* was designed using these principles.

The output coupling transformer was selected to obtain a match between the collectors of the Type 970 silicon transistors and the control phase of a size ten servo motor having a tuned impedance of 220 ohms. Graphical analysis of the Type 970 characteristic curves indicated that a collector to collector impedance of 3200 ohms is optimum for an output of three-quarters of a watt. Recalling the relationship between the reflected impedance between primary and secondary of a transformer, it is seen that the turns ratio should be 3.8 to 1. This circuit operates satisfactorily throughout the range of $-55°C$ to $+125°C$. Cross-over distortion is negligible and power supply stand-by current does not become excessive at elevated temperatures.

Power supply complexity is a direct function of the linearity requirements of the amplifier output *vs.* input curve. Regulated power supplies are called for if extreme linearities are needed. A full wave rectifier, followed by a low resistance filter network will suffice in most applications.

In an effort to reduce size, thought should be given to elimination of the output transformer. This can be done by center tapping the control phase of the servo motor with a resulting degradation of motor performance for a given motor size. Saturation of the motor will not occur. The impedance match between the motor and the collectors must be compatible with the power output requirements and the limiting factors of the transistor.

A much improved impedance match exists between the collectors and the load in a series connected direct feed Class B push-pull stage. See *Fig. 5a.* It can be seen that the entire control phase impedance appears in the output circuit of both transistors and that the *d-c* component of load current furnished by each transistor is in opposite directions through the load thus obviating any saturation difficulties. Cross-over distortion is eliminated in this circuit much the same way as was done previously.

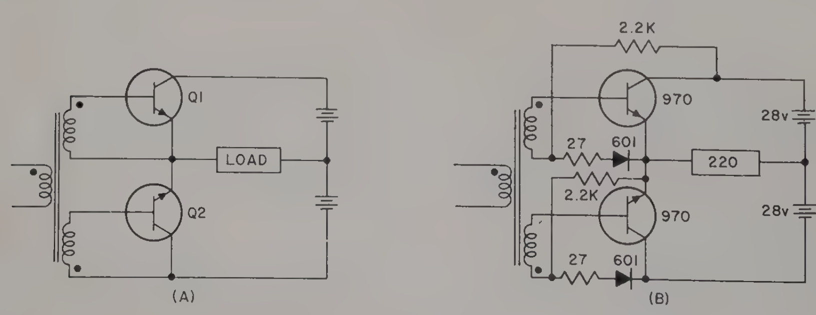


Fig. 5a &5b—Series connected direct feed Class B push pull stage

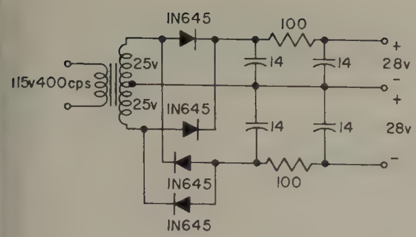


Fig. 6—Center tapped full wave voltage doubler

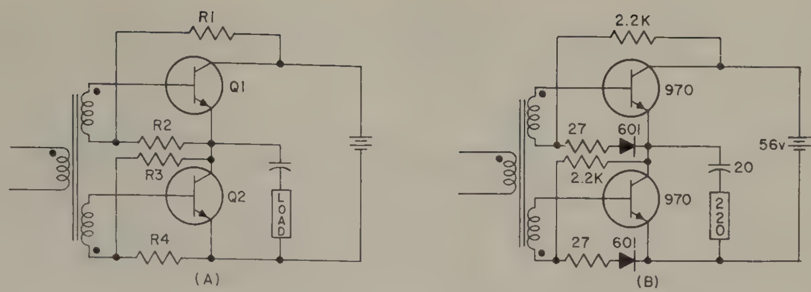


Fig. 7a & 7b—Series connected shunt feed Class B push pull stage

Regulation and power consumption rule out a bleeder network as the source of the two equal voltages required by this arrangement. The most efficient supply is the center tapped full wave voltage doubler circuit of *Fig. 6*. At first glance, this configuration appears to require much more volume than the standard full wave circuit but this is actually not the case. The rectifiers and capacitors can be smaller since each half of the doubler supplies power to only one transistor where in the conventional circuit the one supply feeds both output transistors. The cost of the doubler circuit exceeds that of the standard supply.

The power stage of *Fig. 5b* in conjunction with the supply of *Fig. 6* can also furnish three-quarters of a watt to the motor throughout the range of $-55°C$ to $+125°C$. Unbalanced direct current through the motor control phase does not exceed two milliampere, which is well below the saturation level of the servo motor.

The disadvantages of a center tapped supply, can be overcome by a circuit known as a series connected shunt feed Class B push-pull amplifier, while still maintaining the advantages of the direct feed circuit. Refer to *Fig. 7*. As in the standard circuit it is possible to obtain peak to peak signal output voltages nearly equal to the supply voltage.

Emitter to collector voltage of each transistor should equal one half of the supply voltage for ideal operation of the configuration. Assuming the resistance of the transistors in the quiescent state to be large in comparison to the biasing network resistance it can be seen that the voltage at the junction between the emitter of Q_1 and collector of Q_2 is a function of the ratio between the sum of R_1 and R_2 and the sum of R_3 and R_4. Once the desired voltage at this point is established, the quiescent current through both transistors can be controlled by the relative values of R_1 to R_2 and R_3 to R_4.

Since the collector voltage of each transistor is only one half the supply voltage, temperature stability will be enhanced because of the reduced possibility of thermal runaway. Peak collector voltage is limited to the supply voltage while in the ordinary push-pull circuit collector voltages of nearly twice the supply voltages are experienced. Greater understanding of circuit performance may be obtained from *Fig. 8*.

The voltage gain of this circuit is slightly less than that of the direct feed circuit because of the coupling capacitor. From the gain and phase shift view point the capacitance should be as large as possible. The voltage rating is established by the source voltage. Hence a compromise must be reached between gain and volume when selecting this component.

The configuration of *Fig. 7b* is also capable of delivering three-quarters of a watt to the control phase of the same motor used in the previous circuits throughout the range of $-55°C$ to $+125°C$. As it should be expected, the increase in power supply current during standby operation at elevated temperatures is much less than that observed in the earlier circuits. Since the reactance of the coupling capacitor at the operating frequency of 400 *cps* is small compared to the 220 ohms of the motor control winding there is a gain reduction of only 10% in comparison to the direct feed circuit.

In conclusion, it must be said that each of the transistor servo amplifier output stages discussed here has both advantages and disadvantages. It is up to the engineer to consider each circuit in light of his particular design specifications in order to evolve the optimum amplifier for the intended purpose.

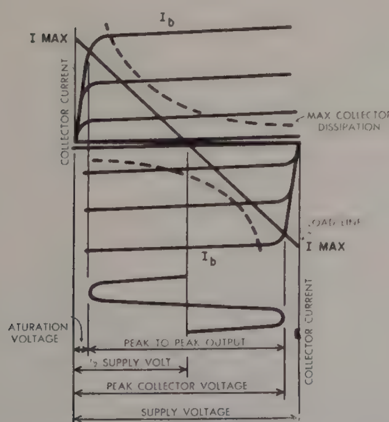


Fig. 8—Graphical presentation of series connected Class B push pull stage

ACKNOWLEDGEMENT

The author wishes to express appreciation to Mr. L. F. Flaczynski for his assistance in the preparation of this paper.

Some Useful Techniques for Transistor Power Gain Measurements

JACOB S. BROWN*

When making power gain measurements on transistors over a wide frequency range it is advantageous to use circuits that can be switched and tuned simply over the frequency range. Circuits such as resistive constant power generators and conjugate tuned constant power loads fall into this category. The principles of several of these circuits are discussed and examples of them are described. A bolometer circuit adapted to a power meter for use at low frequencies is also described. The above mentioned circuits are useful for quick measurements of power gain from ¼ mc. to 40 mc.

INTRODUCTION

THE TECHNIQUES and procedures that are to follow were intended to give quick simple measurement of power gain. The measurements are made at ¼, ½, 1, 2, 4, 7, 10, 20, 30 and 40 *mc*. No claims are made for the uniqueness of the measurement in the sense that it is tranducer power gain or maximum available power gain. It does, however, give measurements which are useful for designing transistor amplifiers.

Method of Measurement

A block diagram of a gain measuring set is shown in *Fig. 1*. It consists of a generator, an impedance matching network from the generator to the transistor input, the transistor and its power supply. The transistor output is matched to a load and some provisions are made for measuring power output. For both input and output matching networks direct or simple impedance calibrations would be desirable. In the case where the power gain of the transistor is too low to measure on the power measuring instrument, then an amplifier can be introduced after the load. The input impedance matching network could take many forms such as transformers, reactance π, *T*, *L* networks, or resistance networks. The simplest to design for the frequency range of operation with a direct calibration is a resistive constant power network.[1]

Components

Constant power resistive generators, such as shown in *Fig. 2*, have been described in the literature by W. F. Chow.[2] For a constant available power, P_{av} with a load R, the voltage is

$$V_g = 2\sqrt{P_{av}}\sqrt{R}\,,$$

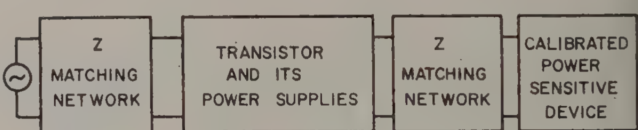
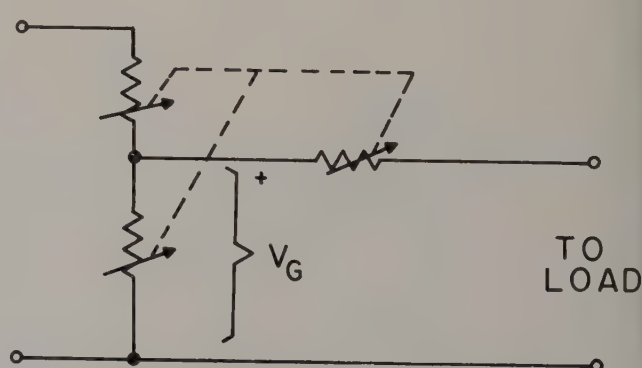


Fig. 1—Block diagram of gain measuring set.



$$P_{AV} = \frac{V_G^{\,2}}{4R}$$

Fig. 2—Constant power resistive generator.

**General Electric Company*
Syracuse, New York

[1]R. L. Pritchard, "High-frequency Power Gain of Junction Transistors" *Proc. of IRE*, vol. 43, pp 1075-1085, Sept. 1955.
[2]W. F. Chow, "Transistor Power Gain Meter" *Tele-Tech and Electronic Industries*, vol. 15, p 104, June 1956.

from which we can calculate the dividing ratio of the input potentiometer. This circuit can be realized for a stepwise changing R by using a taped voltage divider with the series resistance R changed simultaneously. If the input resistance of the divider is 50 ohms then one finds with R varying in steps of 2 to 1 with R changing from a minimum of 25 ohms to a maximum of 3200 ohms that the divider is composed of values ranging from 1½ ohms to 12 ohms. Such a constant power source is adequate for measurements up to a frequency of about 2 *mc*. Above this frequency the inductance of the resistors becomes very troublesome. This was brought home when such a divider was constructed and the dividing ratio at 10 *mc* was found to have no relationship whatsoever to the resistance ratios. Bridge measurements on some sample carbon and boro-carbon film resistors revealed that with leads of ¼ of an inch on each end of the resistor, there was an inductance of about $1\frac{1}{2} \cdot 10^{-8}$ henries in series with the resistors. At first sight this does not appear to be much but at 10 *mc* this results in a reactance of about 1 ohm. Noting that some of the resistors necessary in the divider are near this value and that operation up to 40 *mc* is of interest, some other approach is necessary. One such solution is shown in *Fig. 3*.

The choice of the proper symmetrical attenuator results in convenient values of resistance in the asymmetrical network. By convenient values of resistance is meant resistors of such values that they are neither so high that shunt capacitance is a problem or so low that series inductance is troublesome. The constant power generator used from ¼ *mc* to 10 *mc* is shown in *Fig. 4*. It was designed to deliver 1 microwatt to a matched resistive load when supplied with ½ volt. A 6 db attenuator is available through switching if it is desired to decrease this input power to ¼ microwatt.

To measure power output an ultra high frequency power meter was found convenient because its calibration could be used to read power output with only one mental subtraction. This did, however, require a tuned impedance matching network. For the frequency range up to 7 *mc* a circuit such as that shown in *Fig. 5* was used. Let K_1 be the gain of T_1, K_2 the gain of T_2, "a" the fraction of T_1's output driving T_2, R_B the bolometers resistance and R_L the resistive component of the load. The condition for the same power in the bolometer as in the resistance R_L is

$$\frac{V^2}{R_L} = \frac{(Va\,K_1\,K_2)^2}{R_B}$$

From which

$$a = \frac{\sqrt{R_B/R_L}}{K_1\,K_2}$$

By changing "a" and R_L simultaneously the power sensitivity can be maintained the same while the load resistance is being changed. The circuit used in our measuring set is shown in *Fig. 6*.

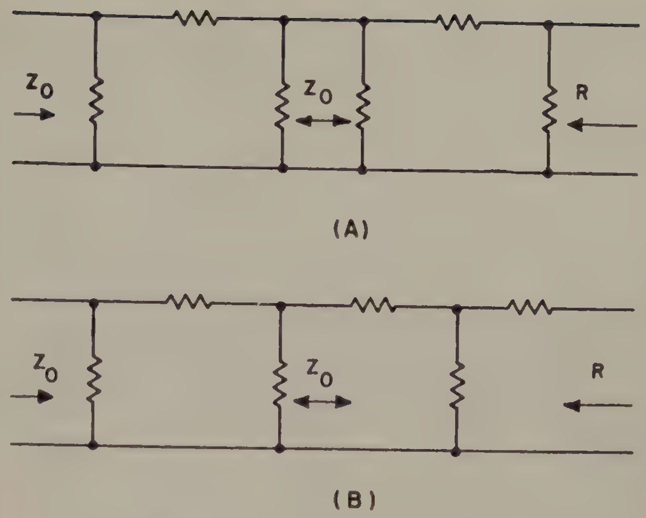


Fig. 3—Symmetrical attenuators.

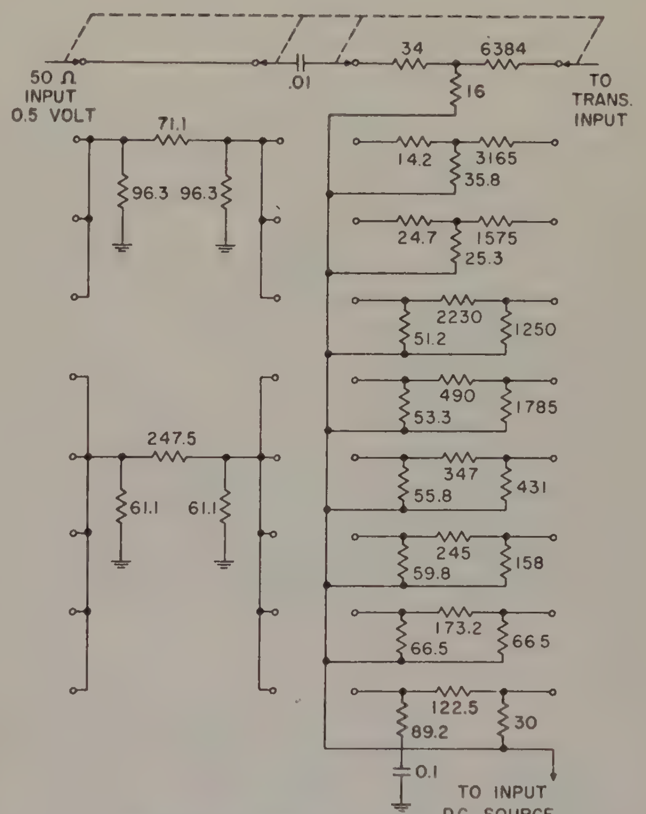


Fig. 4—Constant power generator.

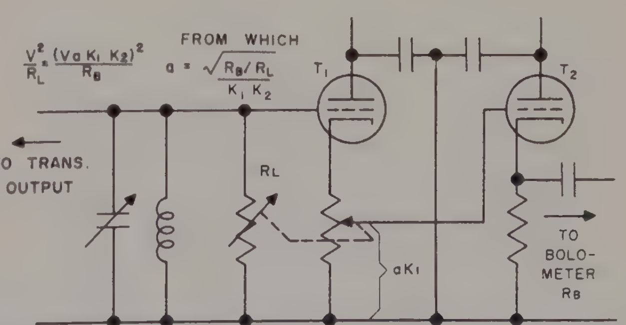


Fig. 5—Matching network from transistor to bolometer.

At the higher frequencies up to 40 *mc* another circuit was found convenient. It is shown in *Fig. 7*. It can be shown that the shunt R_L obtained with such a circuit at resonance assuming the bolometer as a load is

$$R_L = \frac{1}{\omega^2 C_1^2 R_B} + \left(\frac{C_1 + C_2}{C_1}\right)^2 R_B$$

Some errors are introduced by losses in the inductor. They will be discussed later. The variable inductor used had a Q of 300 and an inductance of 13 μh at 10 *mc* and a Q of 150 and an inductance of ¾ μh at 40 *mc*.

Let

$$\frac{1}{\omega^2 C_1^2 R_B} = R_1 \quad \text{and}$$

$$\left(\frac{C_1 + C_2}{C_1}\right)^2 R_B = R_2,$$

where $R_L = R_1 + R_2$. *Figs. 8* and *9* are nomographs that can be used to obtain R_L. The power output is measured with a bolometer and a very high frequency power meter. The power meter used consists of an audio bridge oscillator circuit one arm of which is arranged to be a 200 ohm bolometer. When the power to be measured is supplied to the bolometer, the oscillator changes its output in such a manner that the bolometer resistance remains almost constant. The instrument is calibrated in terms of power to the bolometer necessary to produce a predetermined output. Available instruments have a full scale sensitivity that ranges from 100 microwatts to 10,000 microwatts. The instrument used is also calibrated in *db* with a 1000 microwatt zero reference. When dealing with power outputs smaller than 25 microwatts a distributed amplifier with 200 ohms terminations can be used ahead of the bolometer.

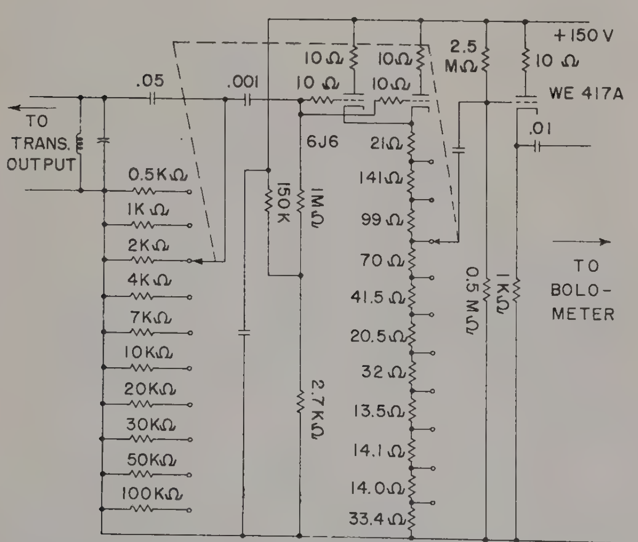


Fig. 6—Measuring apparatus.

Presently available bolometer mounts are for bolometers that present a 50 ohm load to a source. They operate only down to a frequency of 10 *mc*. If the bolometer is to be used as a 200 ohm load and below 10 *mc* then the circuit of *Fig. 10* can be used. The bolometer consists of a 10 milliampere fuse. Remember that it is part of an audio frequency bridge oscillator therefore C_1, C_3, or L_2 cannot be too large or the oscillator will not perform properly. Values of 1000 $\mu\mu f$ for C_1 and C_3 were found to not interfere with the operation of the oscillating bridge circuit. When, however, the signal frequency is lower than 10 mc the reactance

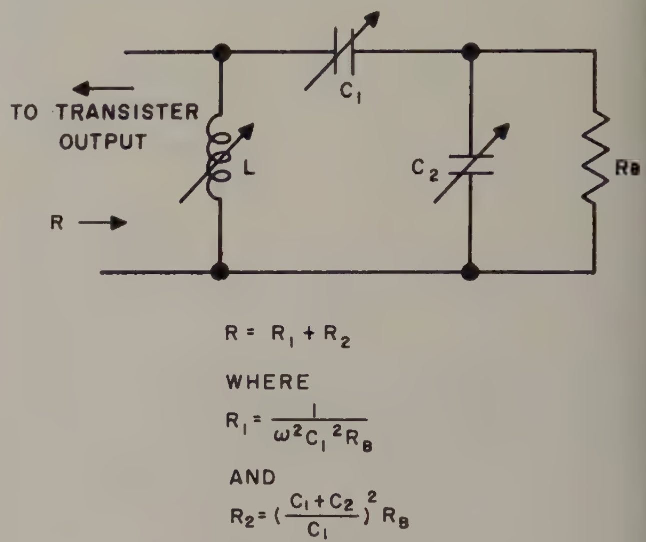


Fig. 7—Measurement set to 40 mc.

of C_1 becomes excessive so L_1 is introduced and series resonated. C_2 and L_2 are parallel resonated to reduce shunting effects on the bolometer. The series inductances and parallel resonant circuits are switched for specific frequencies. It was found that at ¼ and ½ *mc* L_2 was troublesome unless reduced to a value of less than 1 *mh* which would have a reactance of 78½ ohms at 12½ *kc* the frequency of the audio oscillator involved. The adjustment of L_1 C_1 and L_2 C_2 was much more critical at ¼ and ½ *mc* than at the higher frequencies.

It is of interest to note that if the bolometer power meter calibration is checked with a peak reading VTVM then one has to be aware of the purity of signal available from the signal generator. For a 5% second content, readings at two frequencies or different load conditions differing by as much as 0.8 *db* could be obtained with the same VTVM reading. The way around this is to check the bolometer against an *rms* meter. Meters of this type operating over a frequency range of ½ to 40 *mc* are not readily available. A way around this is to use a diode peak VTVM and calibrate one voltage level on it, say ½ volt against the bolometer for the frequencies to be used.

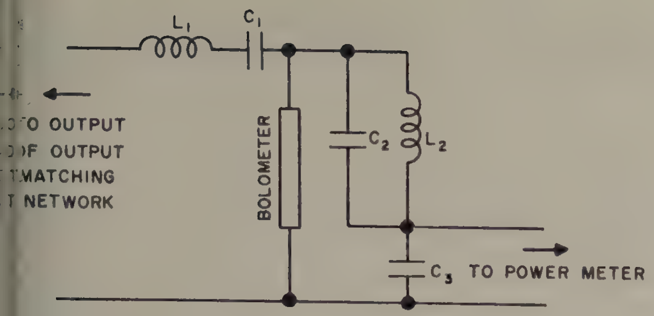


Fig. 10—Use of bolometer below 10 mc.

There are a number of possible sources of error in the scheme described. For example, if the load is resistive and fed from a generator the source resistance of which is changing in two to one steps and if the load impedance has a value just half way between two possible steps of the constant power generator then the power to the load will be 0.11 *db* less than the constant power generator is designed to deliver to a matched load. If the phase angle of the load is 50° and Z of the load matches the resistance of the constant power generator, the load will receive 1.08 *db* less power than the constant power generator is designed to deliver a matched load. If the maximum resistive mismatch to *Z* of 1.33 is considered the load receives .24 *db* less power than the constant power generator is designed to deliver. Another possible source of error is the shunting effect of losses in the output circuit. Consider for example the inductor to be used at the lowest frequency 270 *kc* having a *Q* of 250 and an inductance of 6 *mh*. Its resonant impedance would be 2.5×10^6 ohms. When used with the highest resistance of the constant load generator, it would produce an error of 0.17 *db*. At other frequencies for typical loading conditions of the transistor, expected errors with the constant power load are less than 0.2 *db*, for frequencies up to 7 *mc*. At higher frequencies where the matching network is used, losses in the inductor would be such that with a parallel resistive component of the output impedance of the transistor of 20,000 ohms at 10 *mc* the error should be less than .4 *db*. At 40 *mc*, this same condition would take place with a parallel resistive component of 2400 ohms for the output impedance of the transistor. The variation between these frequencies should be approximately linear.

SUMMARY

Various techniques have been described that can be combined to obtain a useful measurement of transistor power gain. These include a resistive constant power source and constant power load, as well as a π coupling network and nomographs describing its resistance transformation properties. Also included was the adaptation of a microwave power meter to measurements at frequencies as low as ¼ *mc*.

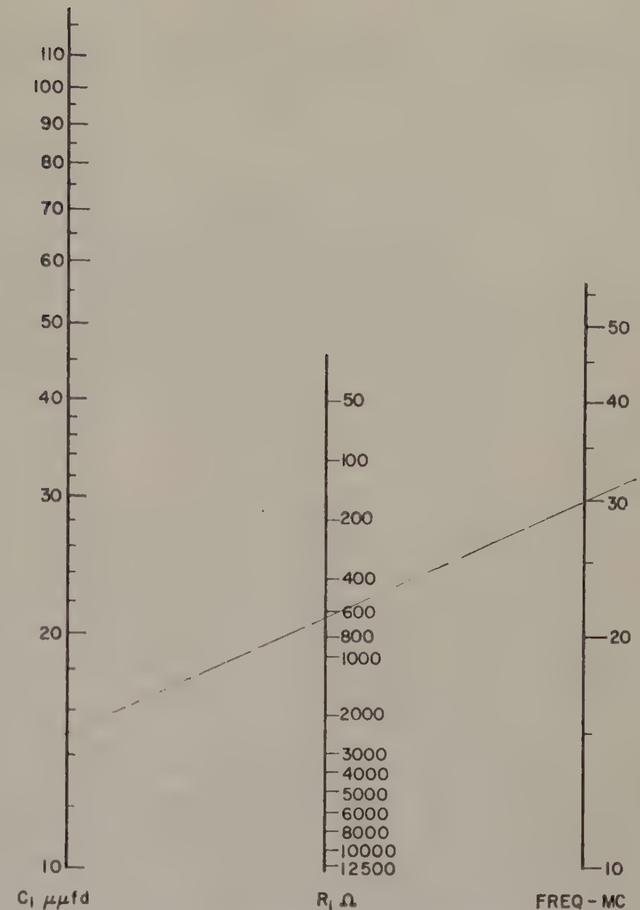


Fig. 8—Nomograph for obtaining R_1.

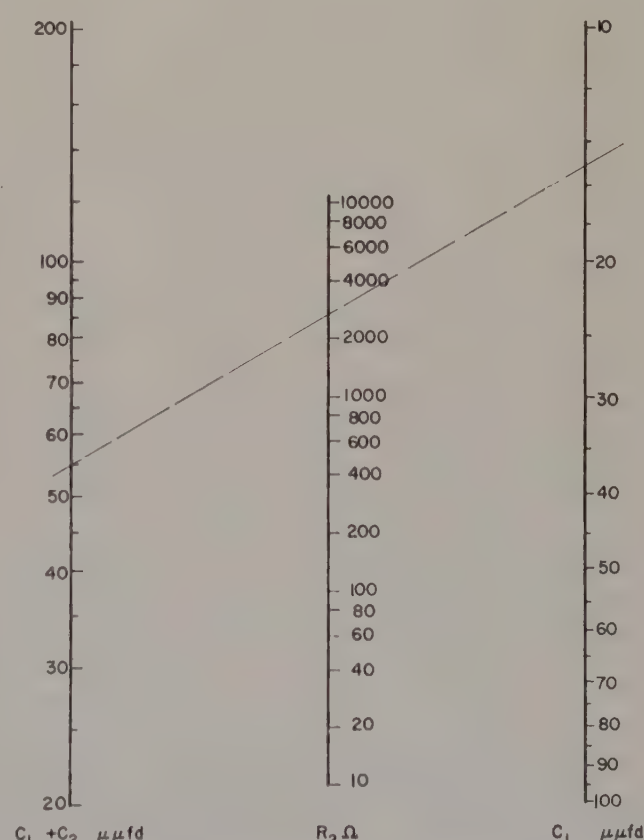


Fig. 9—Nomograph for obtaining R_2.

Design Criteria for Transistor Prosthetic Devices

STEPHEN E. LIPSKY*

The design of transistor amplifiers which are to operate at supply voltages of 1.3 volts imposes several stringent requirements with respct to the proper choice of operating point. Limitations on the maximum signal distortion and minimum acceptable gain further limit the selection of the operating range. Thermal problems are reduced by the restricted temperature range of operation however and circuit economies may thereby be effected. This paper describes some of the specific engineering techniques involved in the design of transistor amplifiers subject to the above limitations and outlines typical circuitry and its method of development.

THE PROPER DESIGN of transistor hearing aids imposes several problems in attempting to achieve very high gain with low battery capacity and subminiature components. In addition, it is necessary to obtain low distortion figures at collector voltages of approximately one volt. For this reason it is difficult to apply the usual temperature stabilization to the emitter circuit and at the same time develop sufficient collector current swing. It is the purpose of this paper to outline some of the techniques peculiar to the design of transistor hearing aid amplifiers used at low battery voltages and to describe typical circuits.

Figure 1 is a photograph of a new Audi-Optic series of hearing aid eyeglasses developed by Electro Acoustic Research Labs. The size requirements for the eyeglass type of hearing aids have dictated the use of *R-C* coupled stages in place of the usual transformer coupled stages found in the larger instruments. The collector voltage drops will then of consequence be greater, and the selection of the proper transistor operating point will be fairly well fixed by the maximum obtainable V_c after the proper load resistance has been determined.

Fig. 1—Audi-Optic hearing aid eyeglasses.

Temperature Stabilization Requirements

Figure 2 is a curve of collector current *vs* gain for a typical *p-n-p* hearing aid transistor. It may be seen that the point of maximum gain occurs at $I_c \cong 0.40$ *ma*. To use this relatively low value of collector current it is necessary to apply only that amount of temperature stabilization required to prevent the saturation current I_{co} from approaching the minimum collector current. Since the maximum collector current swing is dependent upon the gain and the power level at which the stage is to operate, the amount of stabilization necessary for any given stage may be decreased or eliminated entirely if, (a) the value of the I_{co} is predictably low, (b) we limit the maximum temperature to be encountered, and, (c) we operate at small signal levels.

Available transistors for hearing aid manufacture now have excellent saturation current properties compared to earlier units. A typical quantity of units tested had I_{co} values of less than 3 microamperes measured at a collector voltage of 3 volts and at room temperature (27°C). Since the total saturation current is composed of two components; namely the leakage current and the thermally generated current, it is hard to fix the predicted drift of I_{co}. A rather rough rule of thumb is to double the value of I_{co} for every 10°C rise or approximately;

$$(I_{co})_f = I_{co}\ (T_f - T_O)/5 \qquad (1)$$

where: $(I_{co})_f$ is the new value of I_{co} at the final temperature T_f

I_{co} is measured at T_o and at a voltage of 3 volts.

T_o is the initial temperature in degrees centigrade

If the contemplated temperature of operation of the hearing aid is not to exceed 37°C, for a transistor with a measured I_{co} of 3 microamperes at 27°C, the final value of I_{co} will be approximately 6 microamperes.

For a transistor with a beta of 100 the grounded emitter leakage current will rise to 600 microamperes. This places a limit on the minimum current swing,

* Electro Acoustic Research Laboratories

[1]R. F. Shea, "Transistor Audio Amplifiers," John Wiley and Son, N. Y. pp 130-132.

[2]S. E. Lipsky, "Hybrid Hi-Fi Amplifier," *Electronic Design*, May 1, 1957 issue p 78.

in any design, and restricts the temperature operating range. Wherever possible some stabilization should be included.

Developed Circuits

The early stage designs of transistor hearing aids were generally concerned with obtaining optimum gain at low values of operating current. In the eyeglass unit shown, a practical limit on the overall electrical gain figure is approximately 75 *db*. This limit is imposed by the maximum acoustical gain before feedback and the power handling abilities of the input and output transducers.

The stage shown in *Fig. 3* was designed to provide an input impedance of 7000 ohms to minimize the high frequency loading of a nominal 2000 ohm microphone (measured at 400 *cps*). Of the many high impedance transistor circuits available, the common emitter configuration with degeneration in the form of current feedback is the most suitable. By eliminating the usual emitter by-pass capacitor a high input impedance is achieved at the expense of gain, however the stage is linearized with respect to distortion and a small amount of temperature stability is obtained as compared to other circuits, since an added resistor (*R1*) is in the emitter stabilization loop[1]. This method may also be used to provide additional negative inverse feedback from a subsequent stage to adjust the frequency response if desired[2].

The stage shown in *Fig. 3* provides a gain of 23 *db* with input and output impedances of 8000 and 1900 ohms respectively at a collector current of 250 microamperes. The total harmonic distortion is well under 0.75 percent at levels at which this stage is to operate. The transistor *beta* values are in order of 100. The input impedance is fairly constant with different transistors, due to the applied feedback, and the frequency response is far better than that of the available transducers.

Figure 4 is the diagram of another circuit that is best used as an intermediate stage where the extreme constant input impedance is not required. It may be seen that there is no temperature stabilization in the emitter circuit for economy of components and the reasons outlined above. The gain of this stage is 28 *db* and the input and output impedances are 5000 and 2900 ohms respectively. The collector to base voltage of this stage is 0.45 volt which limits the collector voltage excursion to about 300 millivolts of signal swing. The quiescent current is 230 microamperes.

The output stage of a hearing aid that is expressly designed for an eyeglass unit presents several unique problems. A polarizing current of not more than 2 *ma d-c* must flow through the special miniature output transducers, for the proper performance of these units. The output stage therefore must draw a design center quiescent current of about 1.5 *ma* to satisfy this requirement, and in many cases it is necessary to jig select the output transistor to hold accurately to this value.

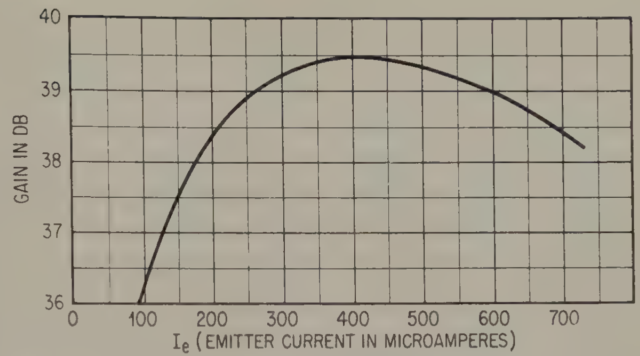


Fig. 2—Collector current vs. gain curve for a typical *p-n-p hearing aid transistor*.

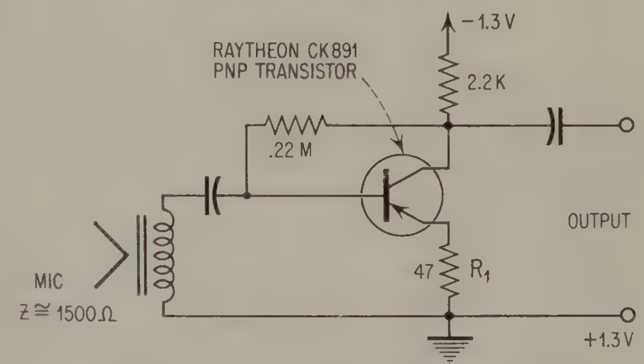


Fig. 3—High impedance input stage with gain of 23 db. Input impedance = or > 8000 ohms. Battery drain = or < 270 microamperes.

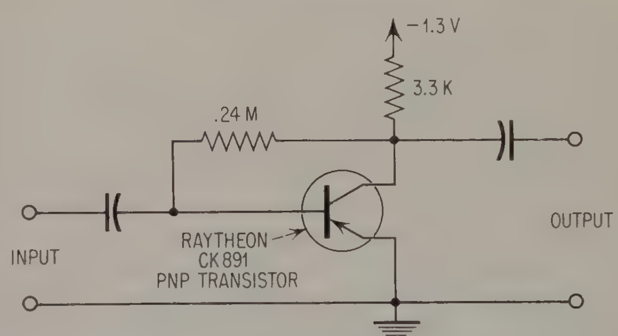


Fig. 4—Intermediate gain stage providing better than 28 db of gain at low distortion levels. Battery drain = or < 230 *uA* at 27° C.

The circuit shown in *Fig. 5* overcomes this and several other difficulties. It may be seen that negative inverse current feedback in the form of emitter degeneration is applied as before. Although this reduces the available collector to base voltage somewhat (0.14 volt), the total harmonic distortion is greatly reduced and values of less than 3 percent may be obtained at reasonably high levels. Another important effect of the degeneration is to improve the overload characteristics of the stage, an important feature in hearing aids where sharp transients must be clipped. The degeneration further aids in the transistor selection and adds some temperature stability. The loss in gain due to the degeneration is compensated for by less matching loss from the previous stage, as a result of the higher input impedance of the output circuit.

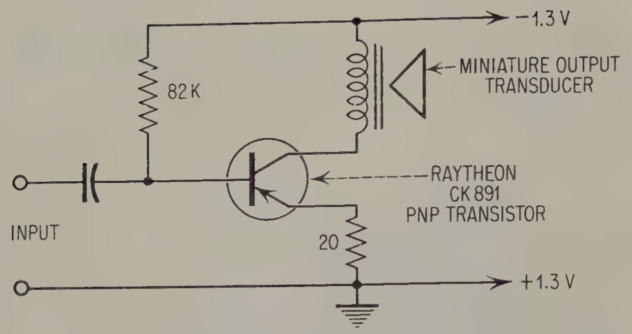


Fig. 5—High level output stage with 21 db of gain. High level distortion in order of 3%.

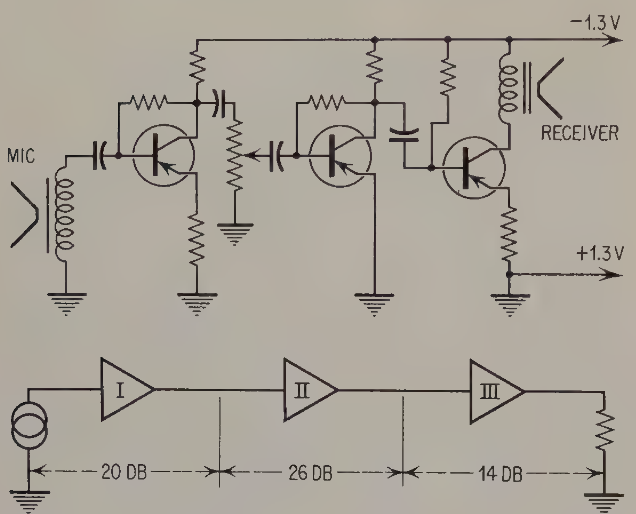


Fig. 6—Three stage transistor hearing aid has gain of 60 db with battery drain of about 2 ma.

Noise Considerations

Early hearing aid transistors were not as satisfactory from a noise consideration point of view as presently available units. It is now possible to obtain quantity shipments of transistors with noise figures of less than 10 *db* with an excellent majority grouping of noise figures at 6 *db*. These figures are measured at a collector voltage of 2.5 volts, a collector current of 500 microamperes and with a generator impedance of 1000 ohms. The noise is measured in a narrow bandwidth centered at 1000 cycles although the integrated noise factor does not differ appreciably from 16 cycles to 20 kilocycles. The value of source impedance of 1000 ohms is favorably matched by the available microphone units of approximately the same impedance.

The theoretical noise voltage E_n, due to the generator resistance alone, may be determined as follows;

$$E_n = \sqrt{4KTR\ (f_2 - f_1)} \qquad (2)$$

where: $K = 1.37 \times 10^{-23}$ Joule/°Kelvin (Boltzman's Constant)
$R = 10^3$ ohms
$T = 300$ °Kelvin
$f_2 = 3500$ cycles/sec
$f_1 = 200$ cycles/sec

The noise as determined from the above is 0.233 microvolt over the frequency range indicated. It should be noted here that the limiting factors in frequency response are the transducers. The equivalent noise input in the first stage, with a transistor having a noise factor of 6 *db*, would be 0.47 microvolt. This value is the minimum voltage necessary to produce a 1 to 1 signal to noise ratio provided that the signal to noise ratio is not deteriorated in later stages.

The signal to noise ratio of overall circuit is difficult to evaluate since the response curve of the microphone is markedly peaked. This peaking is necessary to accentuate the mid-band speech frequencies and to give sufficient output level from the transducer. At normal levels however, the average open circuit microphone voltage is approximately .05 *mv*. With the input noise level as a limit on the overall signal to noise ratio, values of 40 *db* are realizable in most circuits.

Complete Unit

Figure 6 is a complete schematic which is representative of such a 3-stage transistor hearing aid amplifier. The volume control is placed between the first and second stages to maintain a good signal to noise ratio and to prevent early stage overload. This control is isolated from *d-c* to prevent contact noise and has a special taper to give a smooth audiometric response. The total current drain of the circuit is 2.2 *ma* at a battery voltage of 1.3 volts which is supplied by a single Mercury Cell. The overall electrical gain is in excess of 60 *db* and the unit develops less than 3.6 percent total harmonic distortion at rated output, decreasing with level setting.

It is interesting to compare the gains on a stage by stage basis. The first stage develops a loaded voltage gain of 20 *db* which is consistent with design requirements. A higher voltage gain of 26 *db* is provided by the second stage loaded into the high input impedance of the output amplifier, which adds an additional voltage gain of 14 *db*. As compared to the unloaded gain values, the matching loss is in most cases only 3 *db*. This is due, in part, to the higher intermediate impedance levels reducing the usual 6 *db* matching loss. The values of interstage coupling capacitors are reduced somewhat as a result of the higher impedances, typical values ranging from 0.1 to 2 μ*f*.

It is important to evaluate the marginal stability performance of any battery operated device especially where large signal gains are involved. This may be done by inserting a variable resistance in series with the battery and increasing this resistance until low frequency oscillation occurs. With the circuit described no oscillation will occur at any value up to the end-point of the recommended battery supply.

The engineering technique outlined here may be utilized in the design of instruments of higher gain. Although it is possible to achieve greater values of

...n than 60 db by removal of the various feedback ...ps, the distortion figures will be materially deteri...ted. A test of several competitive instruments ...icated that distortion percentages of greater than ... percent were present for gain increase of less than ...b over the three stage instrument.

...Where more gain is required, in cases of severe ...aring impairment, and where acoustical conditions ...rmit, the addition of a fourth stage should be con...ered. In this way the additional gain will not be derived at the expense of good distortion figures. The particular instrument illustrated in *Fig. 1* makes use of this principal. An electrical gain of better than 75 db is obtained in four stages with an increase of only 230 microamperes of battery current as compared to the three stage circuit. The maximum power output of this circuit is greater than the power handling capabilities of many miniature transducers and the total harmonic distortion at these levels is less than 5 percent.

Design Considerations for Transistor Reflex Receivers

ROGER V. FOURNIER*

This paper discusses the use of reflex-amplifier techniques in transistor superheterodyne receivers, and presents practical design information on the construction and performance of such receivers.

IN COMPARISON with a standard five-transistor receiver, the reflex receiver has the advantages of high sensitivity and good frequency response made ...ossible by negative feedback at audio frequencies. ...ts disadvantages, an increased tendency to overload ...n strong signals and a residual volume or "play-...hrough" effect, are minimized by proper circuit de...ign and choice of operating points.

In conventional transistor receiver designs, *i-f* or ...udio-amplifier stages are often sacrificed to reduce ...pace, weight, and cost. The *r-f*, *i-f*, and audio quality ...f the reflex receiver described in this paper compares ...ell with that of receivers having a greater number ...f amplifying elements and components, but its cost, ...pace, and power requirements are much lower than ...hose of the larger receivers.

*Semiconductor Division
...adio Corporation of America
...omerville, New Jersey

Description of Basic Circuit

The reflex receiver shown in *Fig. 1* employs five RCA *p-n-p* alloy-junction transistors as follows: a 2N411 is used as the converter unit, two 2N409's are used as 455-kilocycle *i-f* amplifiers (the second *i-f* transistor also amplifies *a-f* signals), and two 2N407's are used in the Class B push-pull audio-output stage. (The long-lead equivalents of these transistors are, respectively, 2N412 converter, 2N410 *i-f*, and 2N408 audio.) A 1N295 or equivalent diode is used as a second detector.

The receiver operates from a 9-volt supply with a total no-signal current drain of 9 milliamperes, and produces a power output of 200 milliwatts at a distortion of 10 per cent and a "squawk" power output of 300 milliwatts. The sensitivity of the set is 120 microvolts per meter for a 50-milliwatt output and an *a-g-c* Figure of Merit of 60 db (a radiated signal of 500,000 microvolts per meter is used as a reference).

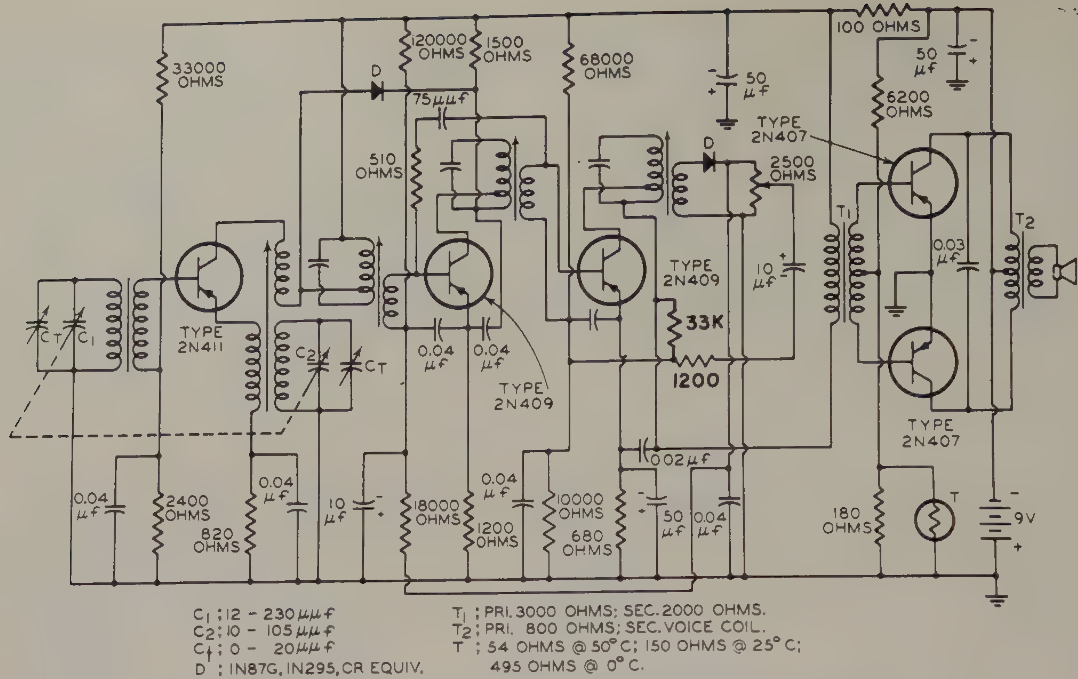


Fig. 1—Schematic diagram for five-transistor reflex receiver.

The antenna circuit shown in *Fig. 1* is derived from a ferrite-loop type of antenna which is highly suitable for receivers of this size. Because the sensitivity of the receiver is proportional to the volume of the ferrite, the largest possible loop consistent with available space in the receiver cabinet should be employed. However, consideration must be given to the orientation of the antenna to avoid unwanted coupling with other circuit elements, particularly the oscillator coil and second-detector circuitry, and the number of ground loops should be kept to a minimum. Harmonic tweet generation can be reduced to acceptable levels by placing critical elements away from the antenna loop, and possibly by shielding.

The converter circuit and associated coil data provide for interchangeability between individual 2N411 transistors without problems of oscillator fall-out at low battery voltage, or improper excitation, regeneration, or blocking at full voltage.

The converter is operated without *a-g-c* to avoid converter cutoff and resultant loss of local oscillation.

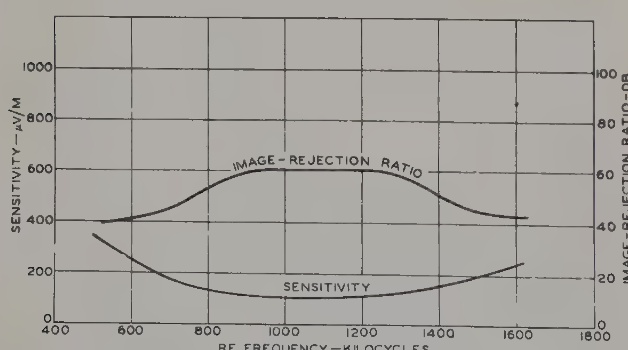


Fig. 2—Image rejection and sensitivity of reflex receiver as functions of frequency.

An overload crystal diode is connected in parallel with the tuned collector circuit of the converter. Its function will be described fully during the discussion of *a-g-c* considerations.

The design of the first and second *i-f* amplifier stages also provides for interchangeability between individual 2N409 transistors. These stages utilize fixed unilateralization, as shown in *Fig. 1*. The *i-f* coil de-

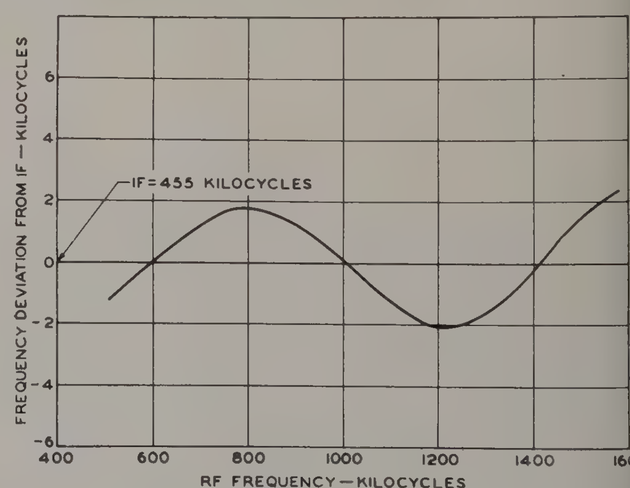


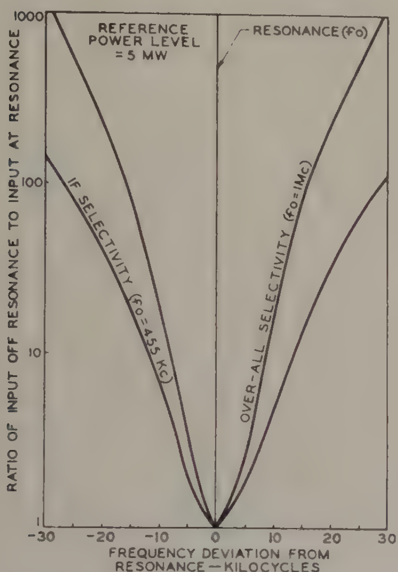
Fig. 3—Tracking curve for transistor reflex receiver.

sign provides maximum gain with excellent *a-c* and d-c stability at temperatures ranging from zero to 55 degrees centigrade.

The second *i-f* stage is a reflex amplifier which amplifies at both intermediate and audio frequencies. The operating point of this stage is critical from the standpoint of (1) overload distortion due to the possible presence at the input of relatively large audio

nals which can excessively shift the quiescent op-ting point, and (2) play-through, *i.e.*, the occur-ice of *a-f* output when the volume control is set at o. The manner in which these disadvantages are olved will be discussed later.

The second detector diode has a dual role. It pro-les an additional degree of temperature compensa-n as well as developing an *a-g-c* voltage for the se of the first *i-f* stage. Although the initial bias ltage on the detector diode is approximately 100 llivolts, the bias increases with temperature so that compensates for the increased collector current in e first i-f transistor caused by increasing ambient nperatures.



ig. 4—Selectivity curves for transistor reflex receiver.

The input of the push-pull Class B audio-output age is transformer-coupled to the reflex amplifier, hich serves as a driving source of excitation. A ermistor is included in the bias network of the out-ut transistor to maintain essentially constant circuit erformance during ambient-temperature variations. hermistor compensation also provides stable opera-ion at rated power output levels in excess of 150 mil-iwatts. The idling current of the output transistors ncreases as the temperature increases and can cause hermal runaway. The thermistor prevents large in-reases in idling current from occurring at elevated emperatures.

The impedance-transformation ratios of the driver nd output transformers are given at the bottom of *ig. 1*. The impedance values for the driver trans-ormer were selected to yield optimum sensitivity at a ow audio-distortion value. The output-transformer urns-ratio was selected to match the RCA high-effi-iency, 2¾-inches 12 ohm speaker, and to provide a rimary impedance which limits the collector current o rated values at high degrees of excitation.

The performance curves shown in *Figs. 2, 3, 4, 5, 6,* nd 7 indicate receiver sensitivity and image rejec-

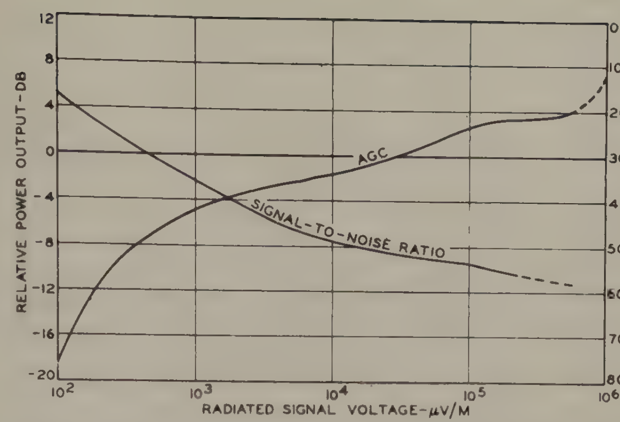


Fig. 5—A-G-C and signal-to-noise ratio as functions of radiated signal voltage.

tion, tracking, selectivity, *a-g-c* and signal-to-noise ratio, distortion, and frequency-response characteristics, respectively. The excellent *a-g-c* characteristics and audio-frequency response of the reflex receiver should be noted.

Residual Volume Effect In Reflex Receivers

Play-through, the occurrence of an audio-frequency output with minimum setting of the volume control, may be a problem in a reflex receiver. This condition may also cause a "minimum-volume effect," in which minimum volume from the receiver is obtained when the volume control is at some setting slightly above zero. At the point of minimum volume, the output signal is very badly distorted because the fundamental frequencies are cancelled out between the normal signal and the out-of-phase play-through signal. Both play-through and, consequently, the minimum-volume effect may be reduced to practical insignificance by proper design.

The play-through effect in a reflex receiver is due to rectification caused by the curvature of the transistor transfer characteristic. Because play-through increases as the input signal is increased, the *a-g-c* system must be designed so that large signals are prevented from appearing at the input of the reflex stage. Because play-through is a function of rectification and, therefore, of the curvature of the transfer characteristic, it is a variable depending on the bias. By proper selection of the bias on the reflex stage, the play-through effect can be made practically negligible. The "minimum-volume effect" is also reduced proportionally because it is caused by the presence of play-through.

The utilization of volume controls which exhibit very little residual resistance at zero setting also helps to minimize the problem of audio output at minimum setting on the volume control. Most potentiometers employed for volume-control purposes on small receivers have a residual resistance of approximately five to ten ohms at zero setting. This resistance may be troublesome, especially when the receiver is tuned to a powerful station in a quiet room. For this reason,

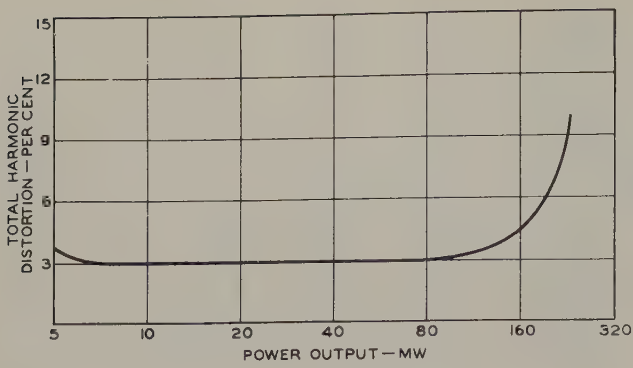


Fig. 6—Total harmonic distortion of reflex receiver as a function of power output.

and because of the play-through effect mentioned above, a reflex stage is usually followed by a power-output stage rather than an audio driver or preamplifier feeding the output stage.

When the volume control of the receiver shown in *Fig. 1* was set at zero, the residual resistance of 3 ohms produced an audio output of less than one milliwatt for a radiated signal of 50,000 microvolts per meter. Listening tests in a quiet room indicated that this level is not objectionable.

A-G-C Considerations In The Reflex Receiver

After the appropriate bias for the reflex stage has been chosen to minimize play-through, a rapid increase in play-through and modulation distortion may be observed when input signals rise above some critical level. This effect occurs whenever the reflexed audio signals become large enough to exceed the bias on the reflex stage and thereby shift its quiescent operating point. In such a case, the *a-g-c* system must be designed to prevent signals of this magnitude from appearing at the base of the reflex stage.

Although the application of *a-g-c* voltage to the converter stage would be desirable to help control overload problems, it would also introduce the possibility of oscillator fall-outs and frequency shifts. Measurements indicate that application of the developed *a-g-c* voltage to the first *i-f* stage does not yield a sufficient a-g-c range to prevent serious play-through and distortion on strong radiated signals. This fact suggests the possibility of applying a fraction of the *a-g-c* voltage to the reflex stage.

However, a good criterion for the "proper" fraction of *a-g-c* is rather elusive. If the fractional *a-g-c* applied to the reflex stage is too small, the rectified *a-f* signal returned to the base from the collector of the reflex stage may exceed the bias on the transistor and cause a shift in operating point with attendant distortion. If the fraction of the *a-g-c* applied to the reflex stage is too large, the *a-g-c* characteristic will tend to reach a maximum output and then fall off with increasing radiated signals. The worst effect of excessive control is the inability of the receiver to deliver full audio output on strong stations even with maximum volume control. Also, in such a case, tuning directly to the carrier of a powerful station produces less output than tuning to one side of the carrier band. As a result, two adjacent tuning positions exist at which maximum volume can be obtained. The reduction in bias caused by the *a-g-c* action also produces a shift in operating point toward the curved region of the transfer characteristic and, consequently, an increase in play-through.

In view of the difficulty of applying fractional *a-g-c* to the reflex stage, a different method of control is employed. In this method, a crystal diode in the collector circuit of the converter is used to prevent overload conditions from occurring on strong signals. In effect, the diode is in parallel with the tuned collector circuit of the converter stage. As shown in *Fig. 1*, the diode is reverse-biased. Under moderate radiated fields, it presents a relatively high impedance to the tuned circuit. With increasing signal strength, the *a-g-c* action decreases the bias on the diode, thereby reducing its impedance. This reduced impedance effectively shunts or "loads down" the tuned collector circuit and causes attenuation of the *i-f* signal level. The effect of this action is to extend the range of the *a-g-c* system so that overloading does not occur when the receiver is tuned to a strong station, and yet full receiver gain can be obtained on weaker signals. The overload diode also helps to maintain a more uniform bandwidth with signal strength.

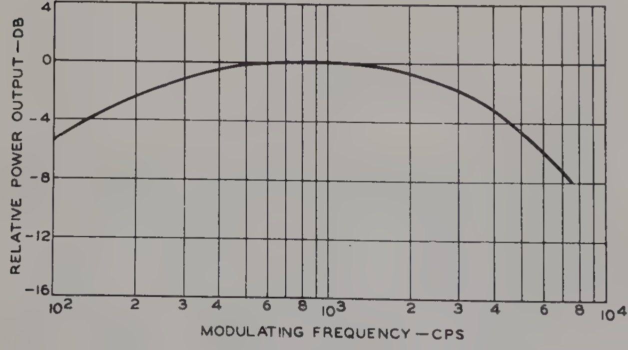


Fig. 7—Frequency-response curve for transistor reflex receiver.

Reflex Gain Stability, Frequency Response, And Distortion

The 33,000-ohm feedback resistor in the reflex stage is used as a degenerative element to prevent excessive variations in over-all gain. The negative feedback also helps to extend the frequency range of the audio response and to reduce distortion and play-through effects. As shown in *Fig. 7*, the 3-db-down points on the audio-response curve occur at 150 and 3000 cycles per second, with a very slow roll-off at both ends of the spectrum. The audio distortion and frequency response of the reflex receiver compare very favorably with that of most receivers of similar size, and, in many instances, its performance is better than that of larger sets.

On Tantalum Capacitors

SAM KASS*

COMPARISON OF CAPACITOR SIZES

TYPE	100 VDC	500 VDC	1000 VDC
Paper Electrostatic	100%	100%	100%
Ceramic Electrostatic	96	250	400
Metalized Paper Electrostatic	65	535	1330
Aluminum Electrolytic	23	30	67
Tantalum Electrolytic	15	25	50

Fig. 1—Comparison of capacitor sizes with the various types of materials used and voltage ratings of the capacitors.

Semiconductor devices lend themselves admirably to miniaturization of equipment. It is not enough however that the semiconductor device itself be miniature in form but that the components associated with the semiconductor be miniature as well. One of the newer devices that has recently taken great strides in miniaturization is the Tantalum Capacitor.

What is tantalum? It is a relatively scarce metal which has replaced aluminum in the manufacture of many types of electrolytic capacitors, resulting in the end product we know as a tantalum capacitor.

The problem of separating the pure tantalum from the rare minerals with which it is always found in combination proved so difficult that it was named after the Greek mythical character, Tantalus, who is represented as standing in water up to his neck with delicious fruits hanging over his head which always elude his grasp when he attempts to reach for them. Certainly all who have sought the elusive tantalum capacitor must appreciate the feelings of the floundering Greek King.

During the century which followed the discovery of the new element in 1803, several methods were worked out which succeeded in the extraction of small quantities of relatively pure tantalum from the ore. But it is only quite recently, after considerable research, that a process was developed for obtaining high-purity tantalum on a production basis. So difficult and costly is this process that the finished metal is valued at about ten times the metal in the natural ore.

The pure tantalum takes the form of a fine powder which is then formed into rough ingots and rolled down into a thin foil. Since tantalum possesses remarkable malleability, it is possible to produce foil as thin as 0.0005″, but the process requires very expensive equipment and a high degree of skill. The cost of tantalum foil may be said to be inversely proportional to its thickness, increasing sharply as the limit of useful tenuity is approached. This property is important because it accounts in large measure for the relative smallness of tantalum foil capacitors.

The chemical inertness of tantalum is its most valuable asset. It is not attacked by hydrochloric or nitric acids or aqua regia under any conditions, and it is able to withstand sulfuric acid of average strength,—even highly concentrated solutions attack it slowly. Its melting point is very high, almost 3000 degrees C.

It may be of interest to know that the first use found for tantalum was in the manufacture of filaments for incandescent light bulbs. These were called tantalum lamps and were patented by Werner von Bolton, a Russian chemist, who was also the first to successfully produce relatively pure tantalum metal in 1905. Unfortunately for von Bolton, the tantalum filament which worked fine on direct current, rapidly crystallized when burned on alternating current. By 1913 the tantalum lamp disappeared from the market.

Balkite rectifiers use a tantalum electrode which is a good conductor when serving as the cathode but quickly forms a nonconducting oxide layer when the current is reversed. The highest testimony paid to the remarkable inertness of tantalum is attested to by its increasing use as a surgical implant in the human body.

There are two important limitations in the use of aluminum for capacitors, these being the existence of copper and iron impurities in aluminum, and the solubility of aluminum oxide in acid solutions no matter how dilute. The impurities (normally present even in 99% pure aluminum), immersed as they are in an electrolyte, react like a battery, causing current to flow, and result in corrosion of the electrodes and a short shelf life. When these impurities appear on the surface of the foil, high leakage currents result accompanied by local heating when voltage is applied.

The ease with which electrolytes dissolve aluminum oxide necessitates the reformation of all such units before they are placed in operation. In circuits where no polarizing voltages are present, an aluminum unit will first increase in capacity as the oxide layer dissolves, then decrease as the electrolyte evaporates.

Tantalum's fitness for use in electrolytic capacitors,—as first shown by A. Guntherschulze in 1926—may be judged by its ability to overcome the inadequacies enumerated above. Tantalum and its most common impurity, columbium, form oxides which are not attacked by electrolytes, resulting in a stable oxide layer requiring no reformation when units are secured from stock after periods of storage. Since dissolution of the dielectric film cannot take place, better sealing techniques for more effective retention of the electrolyte are possible. An additional advantage is the reduced size in comparison with other types. (See *Fig. 1*)

Though acids are desirable as electrolytes because of their excellent conductive ability, the dangerously corrosive effects of such solutions make their use today undesirable. Many Slug capacitors use corrosive solutions because the electrode spacing requires a highly conductive electrolyte. The narrower spacing achieved in Foil construction makes possible the use of somewhat less conductive, but non-corrosive electrolytes. Non-corrosive electrolytes now in common use are lithium chloride, boric acid, and some glycol

*Schweber Electronics

borate solutions. The halide salts are not compatible with easily reactive aluminum oxide. These electrolytes have a very low freezing point and are chiefly responsible for the lower end of the exceptionally wide range of operating temperature rating of the standard tantalum capacitor. (See *Fig. 2*) This range is a freezing minus 55 degrees to a very warm, almost boiling, plus 85 degrees centigrade. The problem of vapor pressure of the electrolyte and the stability of the oxide layer limits the upper temperature range to 85 degrees C. Constructional ingenuity has enabled General Electric to push the upper range to 125 degrees C. This advance is achieved by employing two cases or containers separated by teflon bushings which adds resistance to the vapor pressure and serves as end-seals to protect the electrolyte within the inner case. Also, the 85 degree C working voltage is de-rated by one third.

It might not be amiss at this point to emphasize that the life qualification of the high temperature tantalytic is in excess of 1000 hours continually energized with rated voltage at maximum ambient temperature, that is, 125 degrees C, with no more loss than 20% of the initial capacitance allowed.

At least four distinctly different species of tantalum capacitors are available. Though all use the same tantalum metal, they vary radically in construction. The two most widely used types are best known as "Foil" and "Slug". In spite of considerable differences, they have been used interchangeably in many applications where electrical specifications permitted.

The Slug type consists of a porous tantalum slug sealed into an uninsulated silver case. The slug forms the anode electrode and the silver case serves as the cathode electrode and also doubles as the container for the electrolyte. No mechanical spacers are used. The construction of the Foil type is similar to conventional paper and electrolytic capacitors. Both the anode and cathode electrodes are sheets of tantalum foil separated by a porous spacer material which is thoroughly impregnated with the electrolyte.

It should be remembered that electrolytic capacitors are usually polarized to work when the current flows in one given direction only. The oxide film which serves as the insulating medium will act as a dielectric only as long as the positive

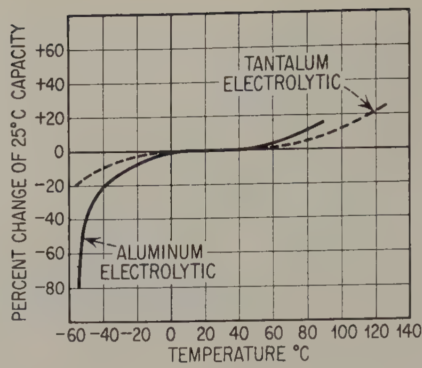


Fig. 2—Characteristic variation of Aluminum and Tantalum Electrolytic capacitance with temperature. Note low temperature characteristic of Aluminum which renders it practically useless at —55° C.

polarity of the anode is maintained. If the polarity is reversed, the capacitor will be destroyed. In view of the foregoing, it is obvious that the Slug type can be used in *d-c* circuits only, because silver is not a film-forming material. On the other hand, the Foil type can be made non-polar because both electrodes are tantalum and both can form dielectric films. Thus each electrode in turn may be positive and then negative without damage to the capacitor because each electrode has its own dielectric film.

A second advantage stems from this duality of Foil type electrodes. Even slight reversals of voltage are a threat to the life of a Slug capacitor. But in the Foil type, the cathode electrode is normally made with a five volt formation of dielectric film thus providing a margin of safety against small voltage reversals. Polarized foil capacitors may be said to be non-polar for voltages under five.

Plain foil capacitors use more tantalum material than the Slug type because both electrodes are of tantalum compared to but one in the Slug. But, by using etched foil design, no more tantalum is needed for a given unit of capacity than is used in a single slug. Etching increases the foil surface area, thus increasing capacity without a corresponding increase in foil material.

Tantalum dielectric films are limited to 150 volts *d-c*. This voltage limitation is imposed by the phenomenon of scintillation or arcing between points on the tantalum through the surrounding electrolyte solution when higher voltages are applied. Foil types are available at 150 volt ratings; Slug types at 125 volts. Higher voltage ratings may be secured by using series arrangements of two or more units. If the *IR* (voltage) drop across each unit is held within 85% of the rated voltage, then no balancing resistors are required.

The contour of the cylindrical Foil type makes it adaptable to subassembly and printed circuitry, and its compatibility with standard resistor and electrostatic capacitor shapes aids compactness of chassis layout. The case of the Foil type is in a "floating" condition and must be insulated by means of mylar sleeving unless the cathode electrode is at common chassis ground. Since the silver case of the Slug type is always negative, it may be mounted directly on the chassis. Within a low voltage and small capacitance compass the Slug construction achieves a smaller sized unit than Foil construction. The production techniques developed for so many years in the making of aluminum foil capacitors are now being directly applied in the manufacture of tantalum foil capacitors.

A third type of construction is known as the Wire type. It is similar to the Slug type except that where one uses a sintered, porous slug of tantalum as the anode electrode, the other uses an etched length or coil of tantalum wire. They are exceedingly small, no more than the size of an oat grain. Their *d-c* voltage rating rarely exceeds 20 at a half µf. or 1 volt at 30 µf. They are used principally in hearing aids.

A fourth, and much more interesting development is the Solid tantalum type in which the "electrolyte" is an inorganic, non-volatile, solid semi-conducting material such as manganese dioxide. Advantages are immediately obvious. Electrolyte evaporation ceases to be a problem, the temperature range drops to —80 degrees C, true hermetic sealing becomes feasible, vibration characteristics are improved, and longer shelf life may be expected. However, the voltage limit of Solid units now available is 60 volts.

The manufacture of tantalum capacitors is a relatively new and exciting field. As such it is expected that the evaluation of experience in field and laboratory will result in continued improvements. No better gauge of the importance of this capacitor development can be cited than the complete engagement in the manufacture of tantalum capacitors by every firm of any consequence in this highly competitive branch of the industry plus a few new ones who jumped in when the initial problems of tantalum fabrication were solved.

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Federal Tool Engineering Co.—Catalog sheet on their complete bench mounted stored energy welder.

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Recent Advances in Luminescence (Cathodoluminescence and Electroluminescence)	IRE Transactions on Component Parts Dec. 1957	Recent developments reviewed also discuss solid-state image display devices and solid-state radiation converters and amplifiers.	H. F. Ivey
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Behavior of Noise Figure in Junction Transistors (Correspondence)	Proceedings of the IRE Feb. 1958	Discussion on modification of noise figure expressions to include effect of partial correction of the emitter and collector noise generators.	W. N. Coffey
The Electrical and Thermal Conductivities, Thermoelectric Power, Hall and Nernst Constants of Amorphous Substances with Electron Conductivities	Soviet-Physics-Technical Physics Vol. 2-No. 1. Translation of Journal of Technical Physics (USSR) by American Institute of Physics	The temperature dependence of various kinetic coefficients for amorphous conductors is calculated.	A. I. Gubanov

CHARACTERISTICS CHART of NEW TRANSISTORS

SEMICONDUCTOR PRODUCTS believes that a tabulation of the new transistors released every four months will be of special interest to its readers. The second of these tabulations, for the period Nov. 1, 1957 to Feb. 28, 1958, is presented here—in type number order and indicating the major characteristics along with the manufacturers of each type. The characteristic symbols are those recommended by the I.R.E.

In the comparatively few years since their inception, transistors have indeed made rapid strides, and it will be noted that in this tabulation, alone, 167 new transistors are included. The characteristics of JETEC registered types are those supplied by the manufacturer of this registered type. This listing is intended merely as a guide; complete specifications, prices, and availability should be obtained direct from the manufacturers.

MANUFACTURERS

(In Order of Code Letters)

MP— Amperex Electronic Corp.
EN— Bendix Aviation Corp.
OG— Bogue Electric Mfg. Co.
THB— British Thomson-Houston Export Co., Ltd.
BS— CBS-Hytron
TP— Clevite Transistor Products, Inc.
EL— Delco Radio Div., General Motors Corp.
EVB— English Electric Valve Co., Ltd.
SEB— Edison Swan Electric Co., Ltd.
THF— French Thomson-Houston Semiconductor Dept.
ECB— General Electric Co., Ltd.
E— General Electric Co., Electronics Div. Semiconductor Prod.
TC— General Transistor Corp.
UG— Hughes Aircraft Co.
IVB— Hivac Ltd.
ND— Industro Transistor Corp.
CTF— Laboratoire Central de Telecommunications
IN— Minneapolis-Honeywell Regulator Co.
OT— Motorola, Inc.
MUL— Mullard Ltd.
NPC— Nucleonics Products Co.
PHI— Philco Corp., Lansdale Tube Co.
PYEB— Pye Industrial Electronics, Ltd.
RAY— Raytheon Mfg. Co.
RCA— Radio Corp. of America, Semiconductor Div.
SIE— Siemens & Halske Aktiengesellschaft
SPR— Sprague Electric Co.
SYL— Sylvania Electric Products Inc.
STCB— Standard Telephone & Cables, Ltd.
TKAD— Suddeutsche Telefon-Apparate-, Kabel und Drahtwerke
TRA— Transitron Electronic Corp.
TFKG— Telefunken Ltd.
TII— Texas Instruments
TUN— Tung-Sol Electric, Inc.
TOK— Tokyo Tsushin Kogyo, Ltd.
WEC— Western Electric Co., Inc.
WEST— Westinghouse Electric Corp.

CHARACTERISTICS CHART of NEW TRANSISTORS

TYPE NO.	USE (See Code Below)	TYPE (See Code Below)	MAT	Max. Ratings @ 25° C				Typical Characteristics	Gain		MFR. (See code at end of chart)
				P_c (mw)	DERATING °C/W	V_{CB}	V_{CE}	$f_{\alpha\beta}$ (mc)	PARAMETER and (condition)	VALUE	
2N115	3	PNP	Ge	50W	1.0	32	32	.20	——	45	AMP
2N235	3	PNP	Ge	25W	2.0	—	40	—	PG	35 db	BEN
2N235A	3	PNP	Ge	25W	2.0	—	40	—	PG	35 db	BEN
2N268A	3, 5	PNP	Ge	14W[7]	1.5	80	60	.300	h_{FE}(2A)	20 min	CTP
2N339	3	NPN(G)	Si	1000	—	55	55	—	PG	30	TII
2N340	3	NPN(G)	Si	1000	—	85	85	—	PG	30	TII
2N341	3	NPN(G)	Si	1000	—	125	85	—	PG	30	TII
2N342	3	NPN(G)	Si	1000	—	60	60	—	PG	30	TII
2N343	3	NPN(G)	Si	1000	—	60	60	—	PG	30	TII
2N350	3	PNP(A)	Ge	45W	1.0	40	30	.30	h_{FE} @ 1A	30	MOT, SYL
2N351	3	PNP(A)	Ge	45W	1.0	40	30	.40	h_{FE} @ 1A	45	MOT, SYL
2N375	3	PNP(A)	Ge	45W	1.0	80	65	.50	h_{FE} @ 1A	65	MOT
2N376	3	PNP(A)	Ge	45W	1.0	40	30	.50	h_{FE} @ 1A	60	MOT
2N394	5	PNP(A)	Ge	150	400	10	10	5.5	h_{FE}(I_c-10ma)	20 min	GE
2N395	5	PNP(A)	Ge	150	400	15	15	7.0	h_{FE}(I_c-10ma)	25 min	GE
2N396	5	PNP(A)	Ge	150	400	20	20	7.0	h_{FE}(I_c-10ma)	30 min	GE
2N397	5	PNP(A)	Ge	150	400	10	10	10	h_{FE}(I_c-10ma)	30 min	GE
2N424	3	(D)	Si	37.5W	—	60	60	—	h_{FE} @ 1A	8	TRA
2N450	5	PNP(A)	Ge	150	400	20	12	6.0	h_{FE}(I_c-10ma)	30 min	GE
2N463	3	PNP(A)	Ge	37W	2.0	60	60	4kc($f_{\alpha e}$)	h_{fe}(I_c-100ma)	83	WEC
2N469[6]	6	PNP	Ge	50	1000	6	—	1.0	AC current gain	10 min	GTC
2N471A	2, 4	NPN(GD)	Si	200	900	30	30	—	h_{fe} @ 1ma	25	TRA
2N474A	2, 4	NPN(GD)	Si	200	900	30	30	—	h_{fe} @ 1ma	50	TRA
2N479A	2, 4	NPN(GD)	Si	200	900	30	30	—	h_{fe} @ 1ma	80	TRA
2N508	2	PNP(A)	Ge	140	250	—	16	3.5	h_{FE}(I_c-20ma)	125	GE
2N509	2, 4, 5	PNP(D)	Ge	200	500	30	—	750	h_{fe} @ 100mc	15.5 db	WEC
2N518	5	PNP(A)	Ge	150	400	45	12	11	h_{FE}(I_c-10ma)	60 min	GE
2N519	5	PNP(A)	Ge	110	—	15	15	1.5	h_{fe}(I_c-1ma)	25	IND
2N520	5	PNP(A)	Ge	110	—	15	12	5.0	h_{fe}(I_c-1ma)	40	IND
2N521	5	PNP(A)	Ge	110	—	15	10	10	h_{fe}(I_c-1ma)	70	IND
2N522	5	PNP(A)	Ge	110	—	15	8	20	h_{fe}(I_c-1ma)	120	IND
2N523	5	PNP(A)	Ge	110	—	15	6	30	h_{fe}(I_c-1ma)	200	IND
2N524	2	PNP(A)	Ge	240	270	45	30	2.0	h_{FE}(I_c-20ma)	35	GE
2N525	2	PNP(A)	Ge	240	270	45	30	2.5	h_{FE}(I_c-20ma)	52	GE
2N526	2	PNP(A)	Ge	240	270	45	30	3.0	h_{FE}(I_c-20ma)	73	GE
2N527	2	PNP(A)	Ge	240	270	45	30	3.3	h_{FE}(I_c-20ma)	91	GE
2N538	3	PNP(A)	Ge	—	2.2	80	—	.24	h_{FE}(I_c-2A)	30	MIN
2N538A	Same as 2N538 with additional specs on power conductance & input resistance										
2N539	3	PNP(A)	Ge	—	2.2	80	—	.30	h_{FE}(I_c-2A)	43	MIN
2N539A	Same as 2N539 with additional specs on power conductance & input resistance										
2N540	3	PNP(A)	Ge	—	2.2	80	—	.38	h_{FE}(I_c-2A)	64	MIN
2N540A	Same as 2N540 with additonal specs on power conductance & input resistance										
2N544	4	PNP(D)	Ge	80	1000	18	—	30	h_{fe}(I_e-1ma)	60	RCA
2N553	3	PNP(A)	Ge	35W	2.0	80	—	$f_{\alpha e}$-20kc	h_{FE}(I_c-.5A)	55	DEL
2N554	3	PNP(A)	Ge	45W	1.0	30	—	.40	h_{FE} @ 1A	45	MOT
2N555	3	PNP(A)	Ge	45W	1.0	40	30	.40	h_{FE} @ 1A	45	MOT

NOTATIONS

Under Use

1—Low power a-f equal to or less than 50 mw
2—Medium power a-f > 50 mw and equal to or less than 500 mw
3—Power > 500 mw
4—r-f/i-f
5—Switching & Computer

Under Type

A—Alloyed
D—Diffused or Drift
G—Grown
M—Microalloy
O—Other
S—Surface Barrier
UNI—Unijunction Transistor

Under fab

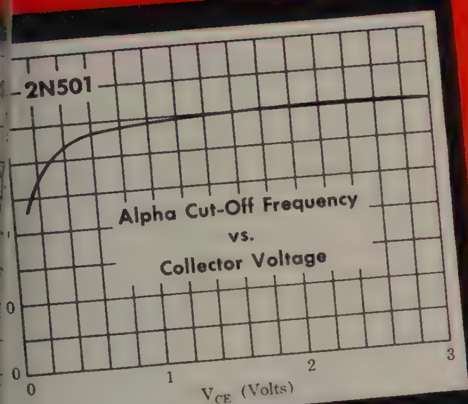
* Maximum Frequency
\# Figure of Merit

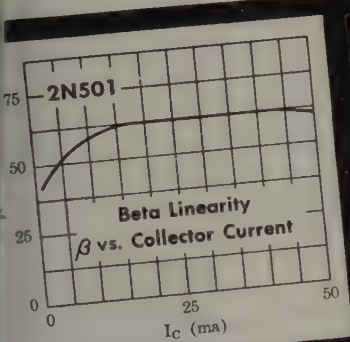
Other

6—Phototransistors
7—@ 70° C
8—I_{CBO} (V_{CB} = —20) = —25 μA
9—Rise time —2 μsec
10—Rise time —3.5 μsec
11—Rise time —6.5 μsec
12—Frequency for fab=1

VHF Transistors! First From PHILCO

New family of Micro Alloy Diffused-base Transistors (MADT)*





- **Rise, Storage, Fall Time in Low mμsec Range**
- **High Oscillator efficiency at 200 mcs**
- **Amplifier gains of 10 db at 200 mcs**

MADT FAMILY APPLICATIONS DATA

TYPE*	f_{max}	Power Gain	Oscillator Efficiency	Class of Use
2N499	250 mcs (min)	10 db at 100 mcs	25% at 100 mcs (min)	oscillator and amplifier to 100 mcs
2N500			25% at 200 mcs (min)	oscillator to 400 mcs
2N501	Ultra high-speed switch typical $t_r = 12$ mμsec; (18 max.); $t_s = 7$ mμsec; (12 max.); $t_f = 4$ mμsec; (10 max.). In circuit with current gain of 10 and voltage turnoff.			
2N502†	500 mcs	10 db at 200 mcs		amplifier to 250 mcs
2N503†		11 db at 100 mcs(min.)		amplifier to 100 mcs
2N504	50 mcs	46 db at 455 KC		high gain IF amplifier

**Available in voltage ratings up to 35V.*
†In JETEC TO-9 Case (widely known as JETEC 30 Case).

Here is a major breakthrough in the frequency barrier . . . a new family of *field-flow* Micro Alloy Diffused-base Transistors. Philco MADT's extend the range of high gain, high frequency amplifiers; high speed computers; high gain, wideband amplifiers and other critical high frequency circuitry.

MADT's are available to various voltage and frequency specifications for design of high performance transistorized equipment through the entire VHF and part of the UHF spectrum. These transistors range in f_{max} from 250 mc to as high as 1000 mc. MADT gains are typically 10 db at 200 mc and greater than 16 db at 100 mc. A low cost general purpose unit is available which will deliver typically 18 db at 50 mc and 32 db at 10 mc.

Make Philco your prime source of information for high frequency transistor applications.

Write to Lansdale Tube Company, Division of Philco Corporation, Lansdale, Pa., Dept. SC-558

*Trademark Philco Corporation for Micro Alloy Diffused-base Transistor circle No. 13 on Reader Service Car

HILCO CORPORATION
ANSDALE TUBE COMPANY DIVISION
LANSDALE, PENNSYLVANIA

TRANSISTOR CENTER U.S.A.

CHARACTERISTICS CHART of NEW TRANSISTORS

TYPE NO.	USE (See Code Below)	TYPE (See Code Below)	MAT	Max. Ratings @ 25° C				Typical Characteristics			MFR. (See code at end of chart)
				P_C (mw)	DERATING °C/W	V_{CB}	V_{CE}	$f_{\alpha\beta}$ (mc)	Gain: PARAMETER and (condition)	Gain: VALUE	
2N556[10]	5	NPN	Ge	100	600	25	—	—	$h_{FE}(I_B\text{-1ma})$	52	SYL
2N557[11]	5	NPN	Ge	100	600	20	—	—	$h_{FE}(I_B\text{-1ma})$	20 min	SYL
2N558[10]	5	NPN	Ge	100	600	15	—	—	$h_{FE}(I_B\text{-1ma})$	60 min	SYL
2N560	3, 5	(D)	Si	600	250	60	60	55	h_{fe} @ 10mc	16 db	WEC
2N563	2	PNP	Ge	150	400	30	25	.80	$h_{FE}(I_c\text{-1ma})$	25	GTC
2N564	2	PNP	Ge	120	500	30	25	.80	$h_{FE}(I_c\text{-1ma})$	25	GTC
2N565	2	PNP	Ge	150	400	30	25	1.0	$h_{FE}(I_c\text{-1ma})$	55	GTC
2N566	2	PNP	Ge	120	500	30	25	1.0	$h_{FE}(I_c\text{-1ma})$	55	GTC
2N567	2	PNP	Ge	150	400	30	25	1.5	$h_{FE}(I_c\text{-1ma})$	100	GTC
2N568	2	PNP	Ge	120	500	30	25	1.5	$h_{FE}(I_c\text{-1ma})$	100	GTC
2N569	2	PNP	Ge	150	400	30	20	2.0	$h_{FE}(I_c\text{-1ma})$	150	GTC
2N570	2	PNP	Ge	120	500	30	20	2.0	$h_{FE}(I_c\text{-1ma})$	150	GTC
2N571	2	PNP	Ge	150	400	25	10	3.0	$h_{FE}(I_c\text{-1ma})$	200	GTC
2N572	2	PNP	Ge	120	500	25	10	3.0	$h_{FE}(I_c\text{-1ma})$	200	GTC
2N574	3	PNP(A)	Ge	—	0.7	60	—	.08	$h_{FE}(I_c\text{-15A})$	12	MIN
2N574A	3	PNP(A)	Ge	—	0.7	80	—	.08	$h_{FE}(I_c\text{-15A})$	12	MIN
2N575	3	PNP(A)	Ge	—	0.7	60	—	.125	$h_{FE}(I_c\text{-30A})$	10	MIN
2N575A	3	PNP(A)	Ge	—	0.7	80	—	.125	$h_{FE}(I_c\text{-30A})$	10	MIN
2N576[9]	—	NPN	Ge	200	375	20	—	—	$h_{FE}(I_c\text{-400ma})$	40	SYL
2N578	5	PNP(A)	Ge	120	—	20	14	5.0	$h_{FE}(I_c\text{-400ma})$	15	RCA
2N579	5	PNP(A)	Ge	120	—	20	14	8.0	$h_{FE}(I_c\text{-400ma})$	30	RCA
2N580	5	PNP(A)	Ge	120	—	20	14	15	$h_{FE}(I_c\text{-400ma})$	45	RCA
2N581	5	PNP(A)	Ge	80	—	18	15	8.0	$h_{FE}(I_c\text{-20ma})$	65	RCA
2N582	5	PNP(A)	Ge	120	—	25	14	18	$h_{FE}(I_c\text{-20ma})$	60	RCA
2N583	5	PNP(A)	Ge	80	—	18	15	8.0	$h_{FE}(I_c\text{-20ma})$	65	RCA
2N584	5	PNP(A)	Ge	120	—	25	14	18	$h_{FE}(I_c\text{-20ma})$	60	RCA
2N585	5	NPN(A)	Ge	120	—	25	24	5.0	$h_{FE}(I_B\text{-1ma})$	40	RCA
2N587[9]	5	NPN	Ge	150	400	40	—	—	$h_{FE}(I_c\text{-200ma})$	20 min	SYL
2N588	4	PNP(MD)	Ge	80	—	20	18	*200min	PG @ 50mc	13	PHI
2N589	3	PNP(A)	Ge	37.5W	2.0	100	100	$f_{\alpha e}$-6kc	h_{FE}(1.5V, 2.5A)	20 min	PHI
2N592	5	PNP	Ge	125	500	20	20	.40	$h_{FE}(I_c\text{-1ma})$	40	GTC
2N593	5	PNP	Ge	125	500	40	30	.60	$h_{FE}(I_c\text{-1ma})$	80	GTC
2N594	5	NPN	Ge	100	600	20	20	1.5	$h_{FE}(I_c\text{-1ma})$	30	GTC
2N595	5	NPN	Ge	100	600	15	15	3.0 min	$h_{FE}(I_c\text{-1ma})$	45	GTC
2N596	5	NPN	Ge	100	600	10	10	5.0 min	$h_{FE}(I_c\text{-1ma})$	60	GTC
2N597	4, 5	PNP(A)	Ge	250	—	30	20	4.5	h_{FE}(1V, 100ma)	40	PHI
2N598	5	PNP(A)	Ge	250	—	30	20	7.5	h_{FE}(1V, 100ma)	100	PHI
2N599	5	PNP(A)	Ge	250	—	30	20	15[12]	h_{FE}(1V, 100ma)	120	PHI
2N602	5	PNP	Ge	120	500	40	—	—	Gain X BW	20	GTC
2N603	5	PNP	Ge	120	500	40	—	—	Gain X BW	40	GTC
2N604	5	PNP	Ge	120	500	40	—	—	Gain X BW	60	GTC
2N605	4	PNP	Ge	120	500	30	—	—	PG @ 2mc	20	GTC
2N606	4	PNP	Ge	120	500	30	—	—	PG @ 2mc	24	GTC
2N607	4	PNP	Ge	120	500	30	—	—	PG @ 2mc	28	GTC

NOTATIONS

Under Use

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3—Power > 500 mw
4—r-f/i-f
5—Switching & Computer

Under Type

A—Alloyed
D—Diffused or Drift
G—Grown
M—Microalloy
O—Other
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UNI—Unijunction Transistor

Under fab

* Maximum Frequency
\# Figure of Merit

Other

6—Phototransistors
7—@ 70° C
8—I_{CBO} (V_{CB} = −20) = −25 μA
9—Rise time −2 μsec
10—Rise time −3.5 μsec
11—Rise time −6.5 μsec
12—Frequency for fab=1

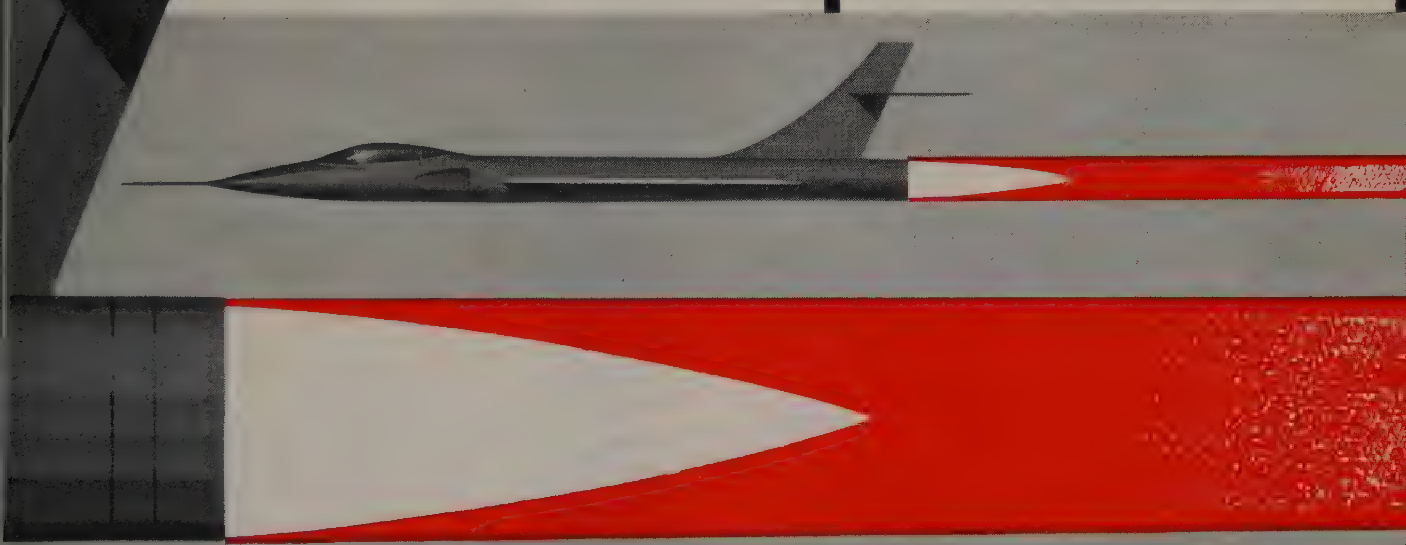
PHILCO

ilicon Transistors

2N495 – 2N496

For outstanding performance at high junction temperatures

- **Excellent performance at Temperatures from −65° C to +140° C**
- **Collector Saturation Voltage of 0.1 Volt or Under**
- **Maximum Frequency of Oscillation in the 15 Megacycle Range**



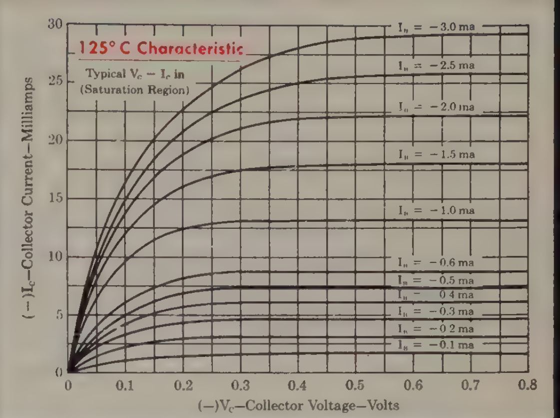
CHARACTERISTICS OF TYPES 2N495 and 2N496

CHARACTERISTIC	CONDITION	TYPICAL VALUE 2N495	TYPICAL VALUE 2N496
Current Amplification Factor, h_{fe}	$V_{CE} = -6$ v, $I_E = 1$ ma	18	
Current Amplification Factor, h_{FE}	$V_{CE} = -0.5$ v, $I_C = -15$ ma		12
Output Capacitance, C_{ob}	$V_{CB} = -6$ v, $I_E = 1$ ma	7 $\mu\mu$f	7 $\mu\mu$f
Maximum Frequency of Oscillation, f_{os} max.	$V_{CB} = -6$ v, $I_E = 1$ ma	15 mc	
Frequency for Beta = 1, f_t*	$V_{CE} = -6$ v, $I_E = 1$ ma, f = 4 mc		15 mc
Cutoff Current, I_{CBO} or I_{EBO}	V_{CB} or $V_{EB} = -10$ v	.001 μa	.001 μa

Maximum Power Dissipation—150 mw

Maximum Collector Voltage 2N495—25 V, 2N496—10 V

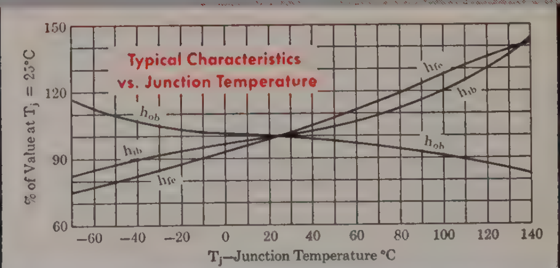
*f_t (the frequency at which beta is unity) is typically 85% of the alpha cutoff frequency.



new Philco PNP Surface Alloy Silicon Transistors permit transistori- of circuits where high ambient temperatures are encountered.

e 2N495 is a general purpose silicon transistor, with excellent per- ce and reliability in amplifier and oscillator applications at frequencies h 15 mc. Units are rated at 150 mw total dissipation with a collector e rating of 25v.

e 2N496 is specifically designed for high speed switching circuits . . . ically over 17 mc. This unit gives the designer the advantages of low ion, low voltage operation and minimum load impedance even at on temperatures as high as 140° C.



CHARACTERISTICS CHART of NEW TRANSISTORS

TYPE NO.	USE (See Code Below)	TYPE (See Code Below)	MAT	Max. Ratings @ 25° C: P_C (mw)	DERAT-ING °C/W	V_{CB}	V_{CE}	Typical Characteristics: $f_{\alpha\beta}$ (mc)	Gain: PARAMETER and (condition)	VALUE	MFR. (See code at end of chart)
2N608	4	PNP	Ge	120	500	30	—	—	PG @ 2mc	32	GTC
2N609[8]	2	PNP(A)	Ge	180	350	—25	—20	1.8	$h_{FE}(I_c$-100ma)	90	WEST
2N610[8]	2	PNP(A)	Ge	180	350	—25	—20	1.5	$h_{FE}(I_c$-100ma)	70	WEST
2N611[8]	2	PNP(A)	Ge	180	350	—25	—20	1.1	$h_{FE}(I_c$-100ma)	50	WEST
2N612[8]	2	PNP(A)	Ge	180	350	—25	—20	.60	$h_{fe}(I_c$-1ma)	40	WEST
2N613[8]	2	PNP(A)	Ge	180	350	—25	—20	.85	$h_{fe}(I_c$-5ma)	60	WEST
2N618	3	PNP(A)	Ge	45W	1.0	80	65	.50	h_{FE} @ 1A	90	MOT
2N619	1	NPN(A)	Si	380	350	50	40	200	h_{fe} @ 5ma	14	RAY
2N620	1	NPN(A)	Si	380	350	50	30	350	h_{fe} @ 5ma	24	RAY
2N621	1	NPN(A)	Si	380	350	50	20	500	h_{fe} @ 5ma	50	RAY
2N622	1	NPN(A)	Si	380	350	50	20	300	h_{fe} @ 5ma	30	RAY
2N623	4	PNP(D)	Ge	40	1000	30	15	200*	$h_{FE}(I_c$-2ma)	35	TII
2N627	3	PNP(A)	Ge	55W	1.0	40	30	.40	h_{FE} @ 10A	20	MOT
2N628	3	PNP(A)	Ge	55W	1.0	60	45	.40	h_{FE} @ 10A	20	MOT
2N629	3	PNP(A)	Ge	55W	1.0	80	60	.40	h_{FE} @ 10A	20	MOT
2N630	3	PNP(A)	Ge	55W	1.0	100	75	.40	h_{FE} @ 10A	20	MOT
2N637	3	PNP	Ge	25W	2.0	—	40	—	$h_{FE}(I_c$-3A)	45	BEN
2N637A	3	PNP	Ge	25W	2.0	—	70	—	$h_{FE}(I_c$-3A)	45	BEN
2N637B	3	PNP	Ge	25W	2.0	—	80	—	$h_{FE}(I_c$-3A)	45	BEN
2N638	3	PNP	Ge	25W	2.0	—	40	—	$h_{FE}(I_c$-3A)	30	BEN
2N638A	3	PNP	Ge	25W	2.0	—	70	—	$h_{FE}(I_c$-3A)	30	BEN
2N638B	3	PNP	Ge	25W	2.0	—	80	—	$h_{FE}(I_c$-3A)	30	BEN
2N639	3	PNP	Ge	25W	2.0	—	40	—	$h_{FE}(I_c$-3A)	22	BEN
2N639A	3	PNP	Ge	25W	2.0	—	70	—	$h_{FE}(I_c$-3A)	22	BEN
2N639B	3	PNP	Ge	25W	2.0	—	80	—	$h_{FE}(I_c$-3A)	22	BEN
3N36	4	NPN(O)	Ge	30	2000	7	6	100	h_{FE}	12.5	GE
3N37	4	NPN(O)	Ge	30	2000	7	6	90 min	h_{FE}	9	GE
B-134	3	PNP	Ge	50W	1.5	—	40	—	$h_{FE}(I_c$-10A)	40	BEN
B-134A	3	PNP	Ge	50W	1.5	—	70	—	$h_{FE}(I_c$-10A)	40	BEN
B-134B	3	PNP	Ge	50W	1.5	—	80	—	$h_{FE}(I_c$-10A)	40	BEN
CK13	4	PNP(A)	Ge	80	750	—	18	3.0	$h_{fe}(I_e$-1ma)	30	RAY
CK14	4	PNP(A)	Ge	80	750	—	15	5.0	$h_{fe}(I_e$-1ma)	60	RAY
CK16	4	PNP(A)	Ge	80	750	—	12	10	$h_{fe}(I_e$-1ma)	80	RAY
CK17	4	PNP(A)	Ge	80	750	—	10	20	$h_{fe}(I_e$-1ma)	140	RAY
CK22	1	PNP(A)	Ge	80	750	—	20	.80	$h_{fe}(I_e$-1ma)	90	RAY
CK25	5	PNP(A)	Ge	80	750	—	20	4.0	$h_{fe}(I_c$-30ma)	30	RAY
CK26	5	PNP(A)	Ge	80	750	—	18	6.0	$H_{FE}(I_c$-40ma)	40	RAY
CK27	5	PNP(A)	Ge	80	750	—	15	11	$H_{FE}(I_c$-55ma)	55	RAY
CK28	5	PNP(A)	Ge	80	750	—	12	17	$H_{FE}(I_c$-80ma)	80	RAY
CK64	1	PNP(A)	Ge	80	750	—	40	.70	$h_{fe}(I_e$-1ma)	22	RAY
CK65	1	PNP(A)	Ge	80	750	—	30	.80	$h_{fe}(I_e$-1ma)	45	RAY
CK66	1	PNP(A)	Ge	80	750	—	20	1.0	$h_{fe}(I_e$-1ma)	90	RAY
CK67	1	PNP(A)	Ge	80	750	—	15	1.2	$h_{fe}(I_e$-1ma)	180	RAY
CTP1127	3	PNP	Ge	14W[7]	1.5	80	60	.300	h_{FE}(2A)	20 min	CTP

NOTATIONS

Under Use

1—Low power a-f equal to or less than 50 mw
2—Medium power a-f > 50 mw and equal to or less than 500 mw
3—Power > 500 mw
4—r-f/i-f
5—Switching & Computer

Under Type

A—Alloyed
D—Diffused or Drift
G—Grown
M—Microalloy
O—Other
S—Surface Barrier
UNI—Unijunction Transistor

Under fab

* Maximum Frequency
\# Figure of Merit

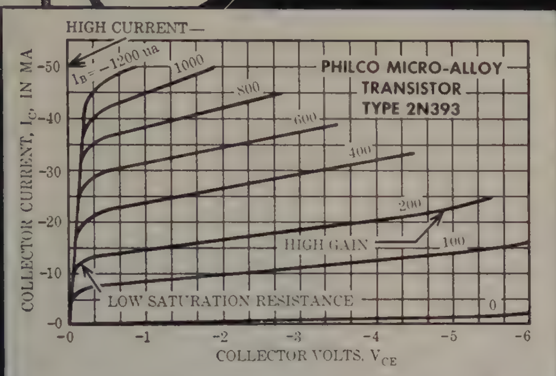
Other

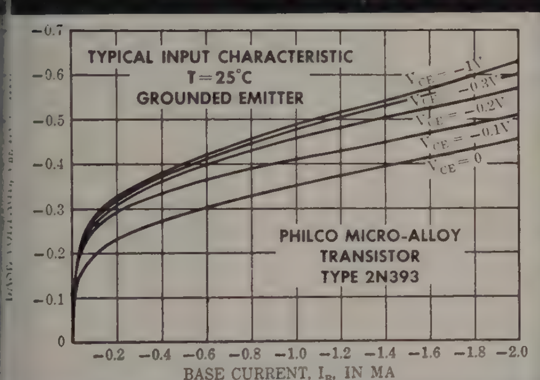
6—Phototransistors
7—@ 70° C
8—I_{CBO} (V_{CB} = —20) = —25 μA
9—Rise time —2 μsec
10—Rise time —3.5 μsec
11—Rise time —6.5 μsec
12—Frequency for fab=1

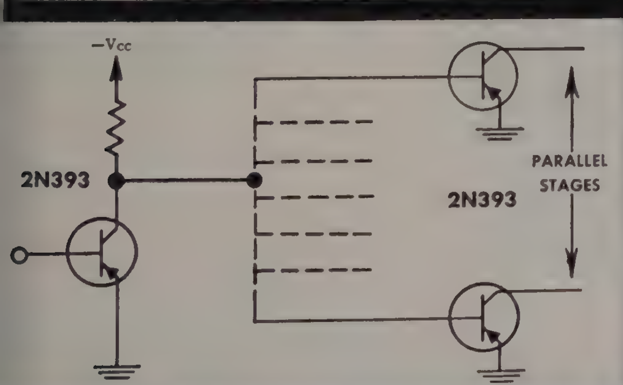
PHILCO

MICRO ALLOY TRANSISTOR 2N393

...Most easily driven, high-speed switching transistor... for modern computer circuitry!







- **Exceptionally Good Life Characteristics, Reliability and Stability**
- **Low Hole Storage**
- **Low Saturation**
- **High Beta at High Currents**

Philco's new 2N393 transistor is exceptionally well suited to the special branching requirements of high speed computer circuitry. Wherever multiple circuits must be driven from a single unit, the new 2N393 significantly outperforms ordinary driven-stage transistors.

The 2N393 combines high gain with excellent high-frequency response at frequencies up to 50 megacycles. Beta linearity is extremely good at currents as high as 50 milliamperes. The new 2N393 micro alloy transistor provides high frequency switching plus low saturation resistance.

This new transistor design is particularly well adapted to direct-coupled logic circuitry. Polarities of the emitter and collector voltages are similar to PNP junction-type transistors.

Make Philco your prime source for complete transistor application information . . .

The 2N393 is also excellent for use in video amplifiers up to one megacycle. For complete specifications and prices on the 2N393, write Dept. S-558

For further information circle No. 15 on Reader Service Card

PHILCO CORPORATION

LANSDALE TUBE COMPANY DIVISION

LANSDALE, PENNSYLVANIA

CHARACTERISTICS CHART of NEW TRANSISTORS

TYPE NO.	USE (See Code Below)	TYPE (See Code Below)	MAT	Max. Ratings @ 25° C				Typical Characteristics			MFR. (See code at end of chart)
				P_C (mw)	DERAT-ING °C/W	V_{CB}	V_{CE}	$f_{\alpha\beta}$ (mc)	Gain		
									PARAMETER and (condition)	VALUE	
CTP1133	3	PNP	Ge	14W[7]	1.5	40	40	$f_{\alpha e}$-20kc	PG @ 2W	30	CTP
CTP1135	3, 5	PNP	Ge	14W[7]	1.5	40	40	.300	h_{FE}(500ma)	40 min	CTP
CTP1136	3, 5	PNP	Ge	14W[7]	1.5	60	60	.300	h_{FE}(500ma)	40 min	CTP
CTP1137	3	PNP	Ge	14W[7]	1.5	40	40	.300	PG @ 2W	34 min	CTP
GA53194	2, 4	PNP(D)	Ge	200	—	30	—	600	h_{fe} @ 100mc	14 db	WEC
GF45017	3, 5	PNP(A)	Ge	825	—	40	40	4.0	$h_{fe}(I_c$-20ma)	215	WEC
GFT2006/30	3	PNP(A)	Ge	10W	5.0	30	20	.40*	PG @ 1.5W	30	TKAD
GFT2006/60	3	PNP(A)	Ge	10W	5.0	60	40	.40*	PG @ 1.5W	30	TKAD
GFT2006/90	3	PNP(A)	Ge	10W	5.0	90	60	.40	PG @ 1.5W	30	TKAD
GFT4012/30	3	PNP(A)	Ge	20W	2.5	30	20	.30*	$h_{FE}(I_c$-400ma)	32	TKAD
GFT4012/60	3	PNP(A)	Ge	20W	2.5	60	40	.30*	$h_{FE}(I_c$-400ma)	32	TKAD
HA5020	5	NPN(A)	Ge	300	150	20	15	4.0	$h_{FE}(I_c$-100ma)	70	HUG
HA5022	5	NPN(A)	Ge	300	150	25	25	4.0	$h_{FE}(I_c$-100ma)	70	HUG
HA5023	5	NPN(A)	Ge	300	150	20	15	8.0	$h_{FE}(I_c$-100ma)	70	HUG
OC30	3	PNP	Ge	3600	14	32	32	.30	——	36	AMP
ST400	3	(D)	Si	60W	—	60	60	—	h_{FE} @ 2A	15	TRA
ST401	3	(D)	Si	60W	—	45	45	—	h_{FE} @ 2A	20	TRA
ST402	3	(D)	Si	50W	—	60	60	—	h_{FE} @ 2A	15	TRA
ST403	3	(D)	Si	50W	—	45	45	—	h_{FE} @ 2A	15	TRA
ST903	2	(G-D)	Si	150	—	30	—	—	h_{FE} @ 1ma	16	TRA
ST904	2	(G-D)	Si	150	—	30	—	—	h_{FE} @ 1ma	31	TRA
ST904A	2	(G-D)	Si	150	—	30	—	—	h_{FE} @ 1ma	60	TRA
ST905	2	(GD)	Si	150	—	30	—	—	h_{FE} @ 1ma	65	TRA
ST910	2	(GD)	Si	150	—	30	—	—	h_{FE} @ 1ma	140	TRA
TR88	5	PNP(A)	Ge	150	—	45	25	1.0	$h_{fe}(I_c$-1ma)	80	IND
TR722	2	PNP(A)	Ge	150	—	45	20	—	$h_{fe}(I_c$-1ma)	22	IND
TR764	5	PNP(A)	Ge	110	—	20	10	25	$h_{fe}(I_c$-1ma)	200	IND

NOTATIONS

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Under Type

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* Maximum Frequency
\# Figure of Merit

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9—Rise time —2 μsec
10—Rise time —3.5 μsec
11—Rise time —6.5 μsec
12—Frequency for fab=1

The following manufacturers have announced that they have just begun supplying the indicated previously registered transistors.

General Electric: 2N332, 2N333, 2N335

Industro: 2N359, 2N360, 2N361, 2N362, 2N363, 2N398, 2N404, 2N422, 2N464, 2N465, 2N466, 2N467, 2N481, 2N482, 2N483, 2N485, 2N486

RCA: 2N356, 2N357, 2N358

Sprague: 2N128, 2N129, 2N393

Sylvania: 2N247, 2N301, 2N301A, 2N312, 2N356, 2N357, 2N358, 2N370, 2N371, 2N372, 2N544

Transitron: 2N117, 2N118, 2N118A, 2N119, 2N389, 2N497, 2N498

New Products

e Voltage Adjuster

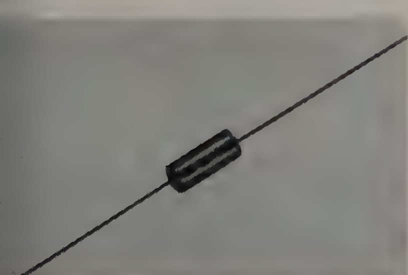
Kepco announces the release of a w Line Voltage Adjuster and pper designed to vary the input ltage for testing the performance electrical and electronic equip- ent. The Line Voltage Adjuster ovides for adjusting and stepping e line voltage from 95 to 135 volts for any fixed input voltage in e range 95 to 135 volts *ac*. The tput capacity of this unit is 3.5 VA for input line voltage above 4 volts. This output capacity de- eases linearly to 3 KVA at an in- it line voltage of 95 volts. The out- it Step Voltage can be adjusted om 0 to 40 volts.

Circle 107 on Reader Service Card

ener Diodes

U. S. Semiconductor makes avail- ble their Semcor axial lead silicon ener diodes which offer all the pace and weight advantages of a ub-miniature package measuring nly .688" x .032", with a lead length f 1½". They are not position-sensi- ive and may be inserted by auto- natic machines on an assembly line asis. An efficient heat-dissipating ath from the Zener diode junction s built-in. This assures better heat ransfer, a conservative power rat- ng of up to 200 *mw* at 25°C, and a igh safety factor in critical applica- ions.

Circle 111 on Reader Service Card



Power Transistors

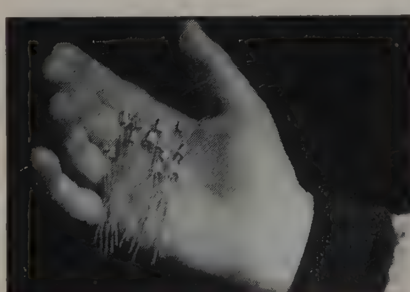
A new series of power transistors is in production at Philco's Transistor Center. Rated from 1 watt to 1.2 watts maximum power dissipation, the family consists of general purpose audio transistors, a 1 *mc* transistor for communication or switching applications and a pulse amplifier. These transistors will find extensive application wherever a transistor of high voltage, current, and power ratings is required. Relays requiring 3 amperes at 40 V may be operated by a transistor in a small studded case with standard basing.

Circle 101 on Reader Service Card

Subminiature Transistors

Raytheon announces their new subminiature type transistors which have a volume only one-fourteenth that of the JETEC-30 package. Four types, CK25, CK26, CK27 and CK28 duplicate the electrical characteristics of the previous Raytheon computer types which were announced in the larger package. Four more types, CK13, CK14, CK16 and CK17 are for general purpose *rf* use, four types, CK64, CK65, CK66 and CK67 are for general purpose audio use. CK22 is a low noise audio amplifier. The transistors are of the fusion-alloy, *p-n-p* type.

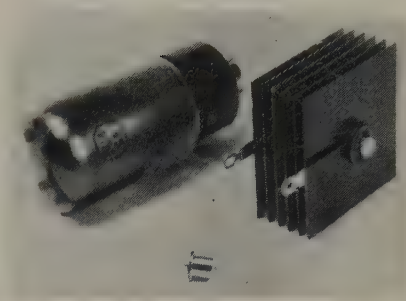
Circle 112 on Reader Service Card



Silicon Rectifier

New "economy-priced" silicon rectifier (foreground) developed by Automatic Manufacturing Division of General Instrument Corporation for high-volume applications in commercial-consumer products, including all types of TV sets, is shown in relation to vacuum tube and selenium rectifiers. Pricing of the PT series is as low as 40 cents to $1.50 each in quantity, depending on the peak voltage required for the application. Rated at 500 milliamperes average rectified current at 100°C, the eight types in the PT series cover a range of 50 to 500 volts. They are designed with convenient pigtail leads for easy mounting in any position.

Circle 109 on Reader Service Card



Transistors and Diodes

Great Eastern Manufacturing Co., manufacturers of electronic components, announces a complete line of transistors and diodes for marketing exclusively through industrial jobbers. All items bear the Gemco tradename.

Circle 190 on Reader Service Card

Switching Transistors

A line of four new PNP medium speed switching transistors having less than a twenty per cent change in h_{FE} and Ico after four-thousand hours storage at 100°C. has been announced by the General Electric Company. The four new germanium transistors have been JETEC type-designated 2N394, 2N395, 2N396 and 2N397. They are designed for use in digital computers and other switching applications where highly stable components are required for maximum overall equipment reliability.

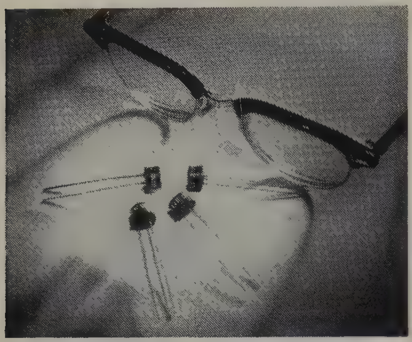
(*Continued on page 56*)

NEW PRODUCTS

(*From page* 55)

The transistors are currently being shipped from stock at G.E.'s transistor warehouse, Buffalo, N. Y.

Circle 108 on Reader Service Card

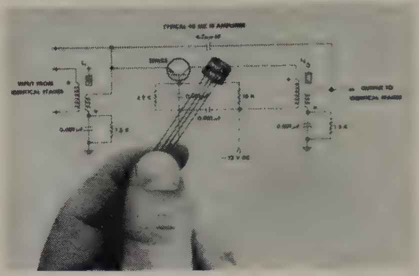


High Frequency Transistors

A new high frequency diffused base germanium transistor was announced recently by Texas Instruments. This device enables manufacturers to realize high gain at high frequencies for television *i-f's*, radio *r-f's*, and very high frequency oscillators. The same transistor, available in a JETEC TO-05 outline package, is especially well suited for use in ultra high speed non-saturated computer applications.

Featuring a 200 megacycle typical maximum frequency of oscillation and a 90 megacycle typical alpha cutoff frequency, this PNP transistor delivers 50 *db* gain at one megacycle and 13 *db* gain at 50 megacycles.

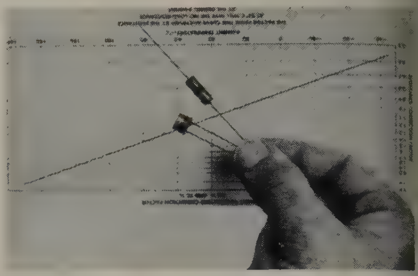
Circle 115 on Reader Service Card



Sensistor

Texas Instruments announces the commercial availability of a new solid state device—the "Sensistor" silicon resistor. The Sensistor has a 0.7%/°C positive temperature coefficient of resistance. There are two configurations of the new "Sensistor" silicon resistor. The TM ¼ is an axial lead molded device which is linearly derated at full load from 100°C to 150°C. The TC ⅛ is encased in a TO-5 round-welded package and is derated linearly at full load from 125°C to 200°C. Both units are immediately available in commercial quantities in standard resistance ratings ranging from 100 to 1,000 ohms at 25°C.

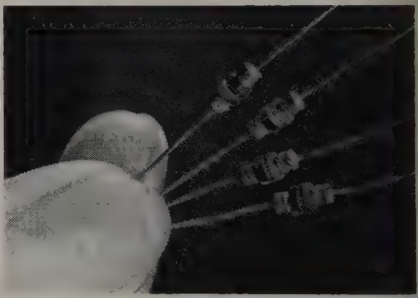
Circle 116 on Reader Service Card



Glass Silicon Rectifiers

Raytheon announces availability of 1N645, 1N646, 1N647 and 1N648 tiny glass silicon rectifiers. These have peak inverse ratings from 225 to 500 volts and are capable of handling 400 milliamperes average forward current at 25°C or 150 milliamperes at 150°C.

Circle 113 on Reader Service Card



Transistor Sockets

Cinch Mfg. Corp., announces their new Universal transistor sockets for use with 10 varieties of transistor bases. Contacts are beryllium copper gold plated. Contacts may be used with either one or two sided 1/16" P.W. boards.

Circle 124 on Reader Service Card

Cleaning Unit

The Model DR 125AH ultrasonic cleaning unit is being offered by Acoustica Associates, Inc. The DR 125 AH can drive two Model AT 200T tanks each of more than ¾ gallon capacity, simultaneously, or one Model AT 500T large tank of 2 gallons capacity. The unit produces 125 watts average, and 500 watts peak power output at 40kc and

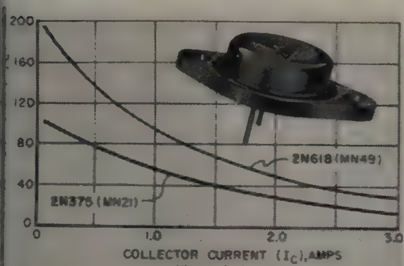
rates from 115V, 50 to 60 cycle
ut. It may be used for the rapid aning of transistors, and numer- other difficult-to-clean items of s nature.

Circle 110 on Reader Service Card

wer Transistor

A new germanium high voltage wer transistor has been announced Motorola's Semiconductor Prod- ts Division. The new transistor, del 2N618, has maximum ratings collector to base voltage of 80 lts, collector current of 3 amps, llector dissipation at 25°C mount- g base temperature 45 watts, col- ctor dissipation at 80°C mounting se temperature 10 watts, and the ounded emitter current gain is ecified as a minimum of 60 and a aximum of 140 at 25°C mounting se temperature.

Circle 114 on Reader Service Card



NPN Silicon Control Rectifier

The GE ZJ39A Silicon Control ectifier is a PNPN semiconductor onsisting of three rectifying junc- ons. Features are, completely static peration—indefinite life with no roblem of wear or noise. High cur- ent ratings: 5 amps continuous, 150 mps peak surge. High PIV's and orward breakover voltages to 300 olts; will operate directly from the ine. Low forward voltage drop— pproximately 1 volt at continuous ating. High temperature operation —full 5 amp rating at 100°C stud emperature. Extremely fast firing and recovery times—approximately 1 microsecond turnon, 3 microsecond turnoff. Large power gains—0.01 watt on control gate switches 1500 watts on anode.

Circle 133 on Reader Service Card

(*Continued on page* 60)

Industry News

Factory sales of transistors in January and February increased considerably over the like months of 1957, the Electronic Industries Association announced recently. Sales of the semiconductor device in February showed a 5 percent increase over January and a 74 percent increase over the February 1957 level of sales.

	1958 Sales (units)	1958 Sales (dollars)	1957 Sales (units)
January	2,955,247	$6,704,383	1,436,000
February	3,106,708	6,806,562	1,785,000
TOTAL	6,061,955	$13,510,945	3,221,000

Transistors rugged enough to still work after being shot from a 12-gauge shotgun into a telephone book were displayed by the General Electric Company as a part of General Electric's exhibit at the Radio Engineering Show in the Coliseum.

Decentralization of the rapidly-expanding Semiconductor-Components division of Texas Instruments Incorporated for the purpose of improving products and customer services and effecting economies, was announced recently by Mark Shepherd, Jr., TI Vice President in charge of the division. Six new product departments, each with complete responsibility for a group of closely related products, have been formed to comprise the division's operations under Cecil Dotson, Manager of Operations. Organized vertically, each department has its own functions of production, product engineering, product marketing planning, production planning, and related activities.

These departments and their managers, with positions formerly held in the S-C division, are: Silicon products—Harry Owens, Germanium products—James McDade, Diodes and rectifiers—J. Rodney Reese, Special germanium devices—Robert Trent, Resistors—Leonard Maguire, and Capacitors—Z. W. Pique.

The 1958 Electron Devices Meeting sponsored by the Professional Group on Electron Devices of the Institute of Radio Engineers will be held October 30 and 31 at the Shoreham Hotel, Washington, D. C.

From the Philco Technological Center Techrep Division, Philadelphia, we are advised that the services of the Philco Technological Center are being made available to Government, Industry and qualified individuals to meet the manpower shortage in technology. The correspondence courses offered and the technical books are practical, up to date and comprehensive. Courses are developed and prepared by a staff of experienced writer-instructors.

...tablishment of the RCA Semiconductor and Ma-...ls Division, responsible for the engineering, ...ufacturing and marketing of semiconductors and ...erials, as well as basic components fabricated ...n them, was announced recently by W. Walter ...ts, Executive Vice President, Electronic Com-...ents, Radio Corporation of America. Mr. Watts ... announced the appointments of Dr. Alan M. ...ver as General Manager of the new Division, and ...liam T. Warrender as General Projects Manager, ...ctronic Components.

... powerful two way radio is now available with a ...y transistorized power supply according to an ...ouncement from Motorola Inc. The new "T-...ver" equipment includes 60 watt radiophones in ... 144-174 *mc.* band and 50 and 100 watt units in ... 25-54 *mc.* band. Four transistors are used in the ... watt radio to replace both the vibrator and the ...amotor.

...osmic ray, meteorite and temperature informa-...n now relayed to earth from the globe-girdling Ex-...rer satellite is being gathered with the help of ...ny tiny silicon transistors (Type 2N328) made by ...ytheon Manufacturing Company.

...A manufacturing plant to keep pace with the ...idly changing semiconductor industry was re-...ntly put in operation by the Westinghouse Electric ...rporation. The plant, near Youngwood, Pa., 35 ...les southeast of Pittsburgh, is the result of the ...mpany's decision to enter the field of semiconductor ...vices primarily for power applications.

Development and start of mass production of ...conomy-priced" miniature silicon power rectifiers ...cousin" to the transistor in the semiconductor ...nily), designed "to break open the virtually un-...pped mass market for semiconductors in con-...mer and commercial products," was announced ...cently by General Instrument Corporation. Pre-...cting that General Instrument's "price break-...rough" would help open up "a $100,000,000 market ...thin two years for silicon rectifiers" ($20,000,000 ...orth of which were produced last year, mainly for ...ilitary and industrial electronic equipment), Board ...hairman Martin H. Benedek announced that the ...ew commercial-consumer devices will sell for as ...w as 40 cents each in quantity, as compared with ...ices of $1.00 to $20.00 for the approximately 180 ...her specialized military-industrial types.

A two-terminal, passive semiconductor component ...aving novel and highly useful characteristics was ...escribed at the annual convention of the Institute of ...adio Engineers in New York. The experimental de-...ice, known as a field effect varistor, was developed ...y R. M. Warner, Jr., H. A. Stone and E. I. Doucette ... Bell Telephone Laboratories. This component has ... constant-current feature which makes it ideally ...ited for a current regulator in circuits where either ...e load or supply voltage vary over wide limits. It ...n also be used as a current limiter or pulse shaper. ...s *ac* impedance is very high, making it useful as a ...upling choke or an *ac* switch.

NEW PRODUCTS

(From page 57)

Medium Frequency Medium Power Transistors

Philco Corporation has a new family of five medium frequency, medium power germanium PNP alloy junction transistors, designed primarily for use in switching circuits. This transistor family features excellent beta linearity, high maximum collector current ratings (400 *ma* for most types), high maximum collector voltage rating (30 V V_{CB}), high beta, and low saturation resistance. Depending on type, these transistors can be used at switching rates ranging from 300 *kc* to 1 *mc*.

Circle 135 on Reader Service Card

Miniature Shift Register

The challenge of volume reduction of electronic circuits has been answered by the announcement by the Sprague Special Products Division of a new micro-miniature shift register assembly. Operational over a temperature range from —55°C to plus 85°C, these epoxy-resin encapsulated units contain one complete shift register stage, diode included, in a ½" x ½" x ⅜" package, less than a tenth of a cubic inch in volume. Designed to operate with a minimum of power, the registers are capable of being driven by transistors or vacuum tubes. Presently available units have maximum operating frequency of 100*kc* and output levels of 6 or 15 volts.

Circle 102 on Reader Service Card

Crystal Growing Accessories

Quartz crucibles, tubing, and distillation apparatus for use in the "growing" of germanium and silicon semiconductors were among the products displayed by Engelhard Industries, Inc., at the 1958 IRE Show in New York. A very pure silicon dioxide quartz must be used to contain germanium and silicon during crystallization, in order to avoid contamination of the materials. The purity required in these semiconductor components is so great that it is measured in parts per billion. Engelhard also showed a radically new gold-coating process called "Atomex." Requiring no electricity or special apparatus, the process imparts a fine sheath of metal by atomic displacement. The coating can be used as anti-corrosion protection and as a solder in certain semiconductor applications.

Circle 103 on Reader Service Card

High Power Transistor

Delco Radio announces a high power, high frequency transistor known as the 2N533. It has solid terminals and the container conforms to the proposed JETEC power transistor outline #1. It is a PNP germanium power transistor intended for military or industrial use where high reliability and optimum specifications are required. Having a low thermal resistance between the junction and mounting base (1°C/watt typical, 2°C/watt, maximum) it is capable of dissipating 12 watts at a mounting base temperature of 70°C.

Circle 105 on Reader Service Card

Solvents

Baker & Adamson announces their new "Electronic-Grade" solvents quality controlled by resistivity measurements. This new technique, coupled with already stringent instrument and chemical analysis methods, makes possible the con-

of impurities in solvents used washing and drying semi-con- tor crystals to a degree surpass- all previous quality standards. ts conducted under plant produc- n conditions indicate direct corre- on between solvent resistivity d percentage of rejects of semi- ductor devices.

Circle 117 on Reader Service Card

wer Tetrode Transistor

Minneapolis-Honeywell Regulator mpany announces a new Power trode H200E. The H200E is a ger- nium pnp alloyed junction tran- tor, characterized by two separate ds to the base, or control layer. uble base construction allows for e of external biasing circuitry, rmitting a high degree of control er the transistor characteristics. e Honeywell Power Tetrode is de- ned to operate from 28-volt col- tor supplies. The current charac- istic permits efficient switching of to 10 amperes. With proper bias- g and input coupling, it is possible obtain uniform current amplifica- n over a wide current range or, if eded, a *rising* amplification vs. llector current characteristic.

Circle 118 on Reader Service Card

ntalum Capacitors

Minitronics Corp. announces their ew tantalum type TQ capacitors hich operate over the temperature nge of —80°C to +85°C with a pacitance variation of only ±10%. he dissipation factor does not ex- ed 0.06 at 120 *cps* and 25°C. The akage current at 25°C is less than 05 microamperes/mfd./volt or 2.0 icroamperes, whichever is greater, easured after five minutes at rated c working voltage applied through 1000 ohm resistor to limit the harging current. The Type TQ is a olarized capacitor to be used where o reversal of potential occurs. The ase is the negative terminal.

Circle 119 on Reader Service Card

ilicon Rectifiers

I.T. & T. announces their type HE- 500 Federal silicon rectifiers. Maxi- num ratings —55°C. to 100°C. ambient. Peak inverse voltage 400 volts. Average d-c rectified current 500 milliamperes. Surge current at 200 millisecond maximum 35 amperes. Recurrent peak current 5 amperes. For medium power supply applications: 1 ampere resistive load at 100°C ambient. 100 to 600 peak inverse volts. Suitable for replacement of tube and metallic type rectifiers with minor circuit modifications.

Circle 123 on Reader Service Card

Wafering Machine

Micromech Mfg. Corp., announces their micromatic precision wafering machine. The fully automatic model WMA is designed specifically for that branch of the electronic industry which processes semiconductor materials used in transistors and diodes. Very thin slicing as well as dicing of any hard, difficult-to-work material, can be accomplished with a high degree of accuracy and speed. Using a 4″, 5″ or 6″ diameter wheel, it produces wafers consistent in thickness and parallelism to within a .0005″ total variation.

Circle 125 on Reader Service Card

General Purpose Transistor

Sperry Semiconductor Division announces their general purpose S-520 transistor designed for medium speed computer circuits and general amplifier operator applications. Features of this device are high temperature operation (140°C), medium frequency operation ($F_{aco} = 1$ *mc*), and 150 *mw* power dissipation with a collector voltage of —30 volts.

Circle 120 on Reader Service Card

Radio and Computer Transistors

Industro features a complete line of PNP germanium radio and computer transistors. One of the services rendered by this company is a brochure containing circuitry for transistor radios, with accompanying coils, transformers, and other component specifications. A transistor interchangeability guide is also made available.

Circle 132 on Reader Service Card

Microwave Crystal Diodes

Kemtron Electron Products, Inc., manufactures a complete line of diodes including microwave selenium crystal diodes. The Kemtron microwave selenium diode plays an important part in the centimeter-wave radars and in many microwave systems where efficient detection or conversion in the kilo-megacycle

range is desired. These complex devices are designed for use as nonlinear circuit elements for frequency conversion, *r-f* and video detection, rectification, modulation and harmonic generation.

Circle 121 on Reader Service Card

Temperature Probes

Veco Lox (Liquid Oxygen) Temperature probes utilizing glass enclosed bead thermistors with resistance values of 50,000 to 500,000 ohms at —195.8°C were exhibited by Veco at the IRE Show. Also exhibited was the Veco Tap-A-Therm, which is a single thermistor with a multiplicity of tapped resistance values. A third new item that aroused considerable enthusiasm and interest was the Veco Compactrol, an ultra sensitive electronic control capable of handling an output of 4½ horsepower with an input signal of only ¼ millionth of one watt!

Circle 134 on Reader Service Card

Tantalum Capacitors

Kemet announces their solid tantalum capacitors which are applicable in transistor amplifiers where low circuit impedance demands high capacitance, low leakage and small size. May be used for coupling, bypass filters and similar applications.

Circle 128 on Reader Service Card

Transistor Stems

Zell Products Corp., announces their JETEC 30 transistor stems which meet the rigid specifications of the military and industry. Bend Test: three 90° bends with one pound weight. Twist Test: three 180° arcs. Leak Test: Mass spectrometer, Dy-Check and Zyglo. Etched Test: To withstand etching in CP4 solutions, nitric hydrofluoric combinations, peroxide solutions, etc. Plate Adherence Test: mandrel test on leads. Corrosion Test: boiling water.

Circle 122 on Reader Service Card

Selenium and Copper Oxide Rectifiers

Bradley Laboratories provides a line of selenium and copper oxide rectifiers, diodes, modulators and arc suppressors. The selenium and copper oxide devices are provided in

variety of basic types of construc-
n. Current ratings of the selenium vices range from microamperes to peres. Voltage ratings for the in-vidual cells are available with -, 33-, and 45-volt *rms* ratings. the copper oxide group half wave ll ratings vary from ¼ to 20 am-res.

Circle 130 on Reader Service Card

ansistor Rise Time sting Unit

Tektronix Inc., features their type /54R transistor rise time unit. The it can be used in all Tektronix scilloscopes with the plug-in fea-re when operated on 50 to 60 cycle e frequency. It supplies a fast-sing pulse and the required supply d bias voltages for measurement transistor rise, fall, delay, and orage times. Rise time of the pulse pplied by the Type 53/54R is less an 5 millimicroseconds, therefore easurement limitations will de-end mainly on the rise time of the scilloscope used.

Circle 131 on Reader Service Card

lapsed Time Indicators

The Simpson Electric Company of hicago, Illinois recently announced e addition of New Elapsed Time ndicators to their large stock of anel instruments. The new Elapsed ime Indicators are available in the amiliar 3½" round (Model 55ET), ½" round shroud (Model 56ET), nd 3½" rectangular (Model 57ET) ase styling to match other Simpson anel instruments.

Circle 191 on Reader Service Card

High Temperature Magnet Wire

Realizing a need for magnet wire for high temperature relay use, Secon Metals Corp., production and development metallurgists, has developed two new types of high-temperature magnet wire. One of these has a specially bonded refractory insulation which is rated for continuous use at 700°F. and for intermittent use up to 800°F. The other wire has a ceramic insulation which is rated for continuous use at 1000°F. Both of these insulations are normally supplied on copper wire. However they may be supplied on other metals or alloys which have better high-temperature characteristics.

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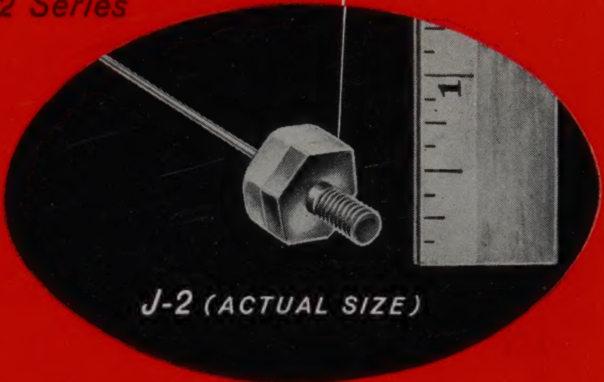

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SPECIFICATIONS

J1 Electrical Ratings

S. T. Type	Max. Peak Inverse Volts	Max. RMS Volts	Current Ratings—Amperes: Max. D. C. Load 100°C	Max. Recurrent Peak 100 C	Surge 4MS Max. 100°C	Jetec No.
10J1	100	70	1.5	10	60	1N1617
20J1	200	140	1.5	10	60	1N1618
30J1	300	210	1.5	10	60	1N1619
40J1	400	280	1.5	10	60	1N1620

For capacitive motor or battery loads, derate d.c. current by 20%

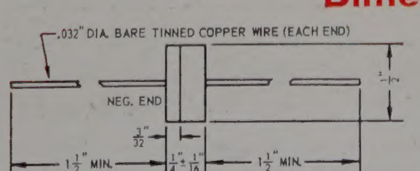
J2 Electrical Ratings*

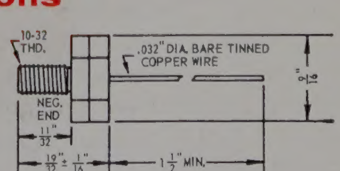
S. T. Type	Max. Peak Inverse Volts	Max. RMS Volts	Current Ratings—Amperes: Max. D. C. Load 100°C	Max. Recurrent Peak 100°C	Surge 4MS Max. 100°C	Jetec No.
10J2	100	70	10	50	80	1N1621
20J2	200	140	10	50	80	1N1622
30J2	300	210	10	50	80	1N1623
40J2	400	280	10	50	80	1N1624

For capacitive motor or battery loads, derate d.c. current by 20%

* with 5" x 5" x 1/16" copper heat sink

Dimensions





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SOLID-ELECTROLYTE TANTALEX CAPACITORS

prompt delivery and complete ratings from SPRAGUE

This solid-electrolyte Tantalex Capacitor (shown 1½ times actual size) is rated at 4.7 µF, 10 volts d-c, and is only ⅛" in diameter by ¼" long.

Sprague, the pioneer in the production of solid-electrolyte tantalum electrolytic capacitors, now offers prompt delivery on production quantities of all standard ratings. New expanded facilities end production delays in your assembly of minified transistor circuits.

Typical of these Tantalex Capacitors is the Type 150D shown above. Its tiny sintered anode is impregnated with a solid, non-corrosive, semi-conductor material which cannot leak under any circumstance. It combines true miniaturization with electrical stability previously unobtainable in an electrolytic capacitor of any type.

Thermal coefficient of these capacitors is sufficiently low and linear so that for the first time a circuit designer can think of an electrolytic in terms of parts per million capacitance change. Nominal value is + 500 ppm/ °C. The capacitor may be used without derating over a range from +85°C to as low as −80°C, a temperature at which no other electrolytic has proved useful.

Solid construction permits the Type 150D to withstand the severe shock and vibration encountered in missile and ballistic applications. Hermetic sealing makes it completely immune to humid atmospheric conditions.

Complete performance data covering the wide range of sizes and ratings are in Engineering Bulletin 3520B, available on letterhead request to the Technical Literature Section, Sprague Electric Company, 467 Marshall Street, North Adams, Massachusetts.

★ ★ ★

Sprague, on request, will provide you with complete application engineering service in the use of Tantalex Capacitors.

DON'T FORGET!—*Sprague also offers prompt delivery on all other types of tantalum capacitors including tubular foil and sintered anode designs.*

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